THE UNIVERSITY OF ALBERTA

A SOLID STATE CONTROL SYSTEM FOR CONTINUOUS WAVE MAGNETRONS

A Thesis

Submitted to the Faculty of Graduate Studies
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DEPARTMENT OF ELECTRICAL ENGINEERING

by

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ABSTRACT

Two primary control systems are used in power supplies for continuous wave magnetrons of the fixed field type. A heater voltage control system is necessary to neutralize the heating effect of electron back-bombardment of the cathode which changes as the power output of the tube is changed. A control system is included to set and regulate the average value of the anode current. The R.F. power output of the tube is proportional to the average value of the anode current.

Present power supplies use saturable core reactors in both control systems for this type of magnetron. The advantages of using silicon controlled rectifiers and transistorized control circuitry in the power supply control systems are investigated in this thesis.

It is found that the solid state control systems offer faster transient response and better regulation to line voltage changes. Because the anode current is controlled by a low power, low voltage signal, programmed operation is possible. Logic type control systems can also be incorporated to back off the magnetron.

In some microwave heating systems where more than one tube is used, it is necessary to prevent cross coupling between individual tubes and to conserve power when loading is intermittent or temporarily absent. This can, in principle, be simply achieved with the solid state control system.

A brief discussion is included at the end of this thesis of other means of controlling the continuous wave magnetron. Reference is made to the cold cathode developed by Raytheon, and the variable magnetic field magnetron developed by Litton.

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1. INTRODUCTION

1.1 Microwave Power

The principles involved with microwave heating are not new, they have been known for over 20 years. It is only recently, however, that the power output and efficiency of tubes has become high enough to warrant their use in industrial processing systems.

The military is largely responsible for the development of high power microwave tubes. The invention of radar during World War II provided the stimulus. After the war, research and development continued, sponsored mainly by the military, towards increasing both the output power and operating frequency.

Recently, with the decline in military purchases, manufacturers have turned their efforts towards developing tubes suitable for industrial use. Military tubes are not especially suited to industrial applications, mainly because of their high cost and unnecessary ruggedness.

At present, three types of tubes are being manufactured. They are the magnetron, the amplitron, and the klystron. The first two are crossed field tubes, the latter being a linear beam tube.

Klystrons are presently capable of producing 500 kilowatts of continuous power at efficiencies between 40% and 60%.[1] Magnetrons are capable of higher efficiencies; one tube presently operates at an efficiency of 80%.[1]

Multiple tube systems are also feasible; up to eight klystrons have been coupled to a common waveguide.[2]

1.2 Microwave Processing

Ordinary methods of heating materials depend on the application of heat to the surface of the material and conduction of the heat to the inside. In order to achieve rapid heating, it is necessary to increase the surface temperature which, if carried above certain limits, can be damaging to the material.

Microwave heating is a volume type of heating. Molecules, inside the material to be heated, are excited by high frequency fields, and the resulting molecular friction causes heat generation. Very rapid heating occurs because heat is generated throughout the material, instead of being applied at the surface as in conventional methods. Precise control is maintained because there are no thermal time lags in a microwave system.

Microwave energy is presently being used to heat, dry, and cure dielectric materials. Microwave energy is particularly suited to drying applications because of a phenomenon known as selective coupling.

If a homogeneous material is heated in a cavity which has a multiple number of excited modes, the heating will be relatively uniform. If, however, the material is not homogeneous, the energy is coupled into the more lossy regions. In drying applications the moist regions are the more lossy regions. Materials emerge from microwave drying with a relatively uniform moisture content.

Selective coupling has been found useful in the manufacture of magnetic tape.[3] In the conventional production of magnetic tape, iron oxide is deposited on sheets of mylar in a binder system that contains a solvent which must be evaporated. Removal of the solvent cures the binding system and makes

the iron oxide particles adhere uniformly to the mylar base. Because microwave energy is not coupled to the iron oxide or the mylar base, the solvent can be evaporated rapidly without heat damage to the tape. Curing times are reduced from hours to seconds.

Microwave power systems have found limited use in rather specialized applications. In one experiment[4] microwave energy was used to ionize gas inside a resonant cavity. The gas ions, upon leaving the cavity, recombined to produce an exceedingly hot plasma flame (3000° Centigrade). In another experiment[4] rock was heated internally by the application of microwave energy. The resulting thermal expansion was sufficient to fracture the rock. Such a system is now being tested as a possible replacement for the conventional jack hammer.

Another field which is also being investigated is the wireless transmission of power at microwave frequencies. The Raytheon hover craft is an example of development in this field.[5]

1.3 Microwave Power Systems

A typical microwave power system contains four basic components; a microwave generator, its power supply which is also used to control the generator, some form of applicator, and a wave guide to couple the applicator to the generator.

Intensive research is being conducted to improve all the basic components. Particular emphasis is being placed on increasing the efficiency and output power of generators and on reducing the losses in applicator structures. With future developments, the transmission of large quantities of energy

at microwave frequencies may become economically feasible.

Difficulties in obtaining a uniform distribution of energy within material being heated has prompted research into multimode cavity design. Mode stirring devices such as fans, metallic reflectors, and pneumatically positioned walls are recent developments.

Power supplies are also subject to consideration. A solid state control system is now possible because of recent developments in solid state devices. The silicon controlled rectifier, in particular, can be used to control AC power in the megawatt range.[6]

This thesis investigates the use of solid state control systems in a power supply for low power (one and two kilowatt) continuous wave magnetrons. Emphasis is placed on the advantages of such a system as compared to present saturable core reactor supplies for the fixed field type magnetrons.

1.4 General Description of a Magnetron[7]

The continuous wave magnetron is, in principle, a special form of vacuum diode. It consists of a massive cylindrical anode with a central space through which the cathode passes. In the body of the anode are a number of cavities arranged parallel with the axis of the magnetron. Each cavity "communicates" with the central space by means of a slot.

When the cathode is heated, electrons travel radially from the cathode to the surface of the anode which is held at a positive voltage with respect to the cathode. A strong magnetic field is set up parallel with the axis of the tube and therefore at right angles to the electric field. The electrons emitted by the cathode travel towards the anode along a cycloidal

path in the course of which they pass across some of the slots.

Each of the cavities in the anode is a resonant circuit. The surrounding metal walls form the inductive component, and the slot forms the capacitive component. During each r.f. cycle, in each cavity, current flows along the surface of the metal from one lip of the slot to the opposite lip and then reverses and travels back. This process continues at the resonant frequency of the cavity resonator. The potential of each lip of the slot therefore oscillates above and below a mean value. Oscillating fields are thus set up in the region of the slot.

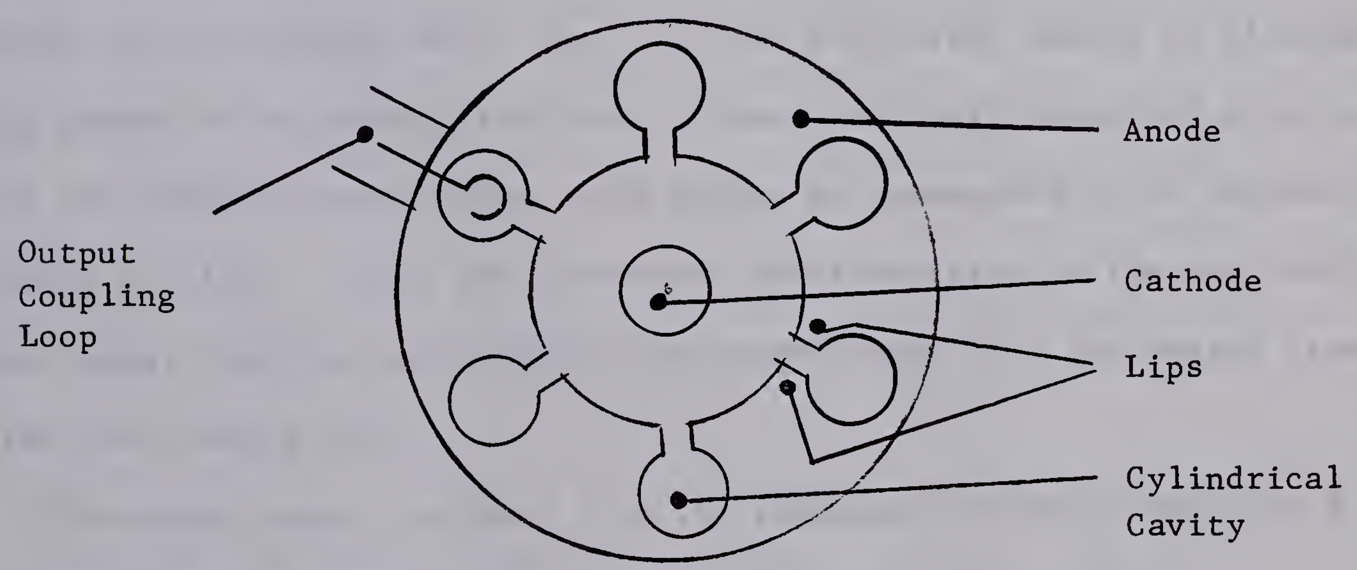


Fig. 1.1 Top View of a Magnetron

An electron, emitted by the cathode, and acted upon by the electrical and magnetic fields, will follow a path in which it passes across one or more slots. If a particular electron happens to enter the oscillating field

at an instant when the direction of this field opposes the force due to the magnetic field, the electron will be retarded and will therefore lose some of its kinetic energy. The energy lost by the electron is imparted to the oscillating field and thus helps to maintain the oscillations. Such an electron usually passes across several slots and imparts energy to them, before hitting the anode with a low energy impact. If, on the other hand, a particular electron enters the oscillating electric field at an instant when the direction of this field is the same as the direction of force due to the magnetic field, the electron will be accelerated and will receive energy from the oscillating field. This tends to damp the oscillations. Such unfavourable electrons are still further deflected by the magnetic field and return rapidly to the cathode, which they hit with sufficient energy to dislodge a large number of secondary electrons. These secondary electrons will, in turn, reach the neighbourhood of the anode either at favourable or at unfavourable instants of time. Since the favourable electrons stay in the oscillating field longer than the unfavourable electrons, there is a net energy transfer to the oscillating field.

The energy output is taken from the resonator system by means of a slot or a coupling loop penetrating into one or two of the cavities.

Summarizing, the magnetron can be considered as a converter in which energy from a DC or low frequency source is converted into microwave energy. Part of the energy applied to the magnetron is lost in the collision of the favourable electrons with the anode. The efficiency of one and two kilowatt magnetrons lies between 50% and 65%.

1.5 Power Supply Considerations

Heater Control System

A magnetron power supply serves two purposes; it provides the necessary heater power to maintain a constant cathode temperature, and it provides the necessary bias voltage and current for the anode. Most magnetrons (the Amperex DX 260 is an exception) require a reduction of heater voltage once the magnetron is operating. This is necessary because heat generated at the cathode, by electron back bombardment, would otherwise increase the cathode temperature.

With the type of cathode used in c.w. magnetrons it is essential that the cathode be kept at its proper operating temperature for maximum tube life. The cathodes used in Philips magnetrons, for example, are of the dispenser type. The body of this type of cathode consists of porous sintered tungsten impregnated with a substance which forms the stock for the emitting layer at the surface of the tungsten body. The operating temperature, the size of the pores, and the composition of the impregnant are such that at the surface a state of equilibrium is obtained between the amount of emissive material lost, and the amount which is replenished by thermal diffusion from the internal storage pores.

Underheating or overheating of the cathode may disturb the equilibrium. With overheating there is a risk of too much emissive material being lost. With underheating the stored impregnant does not diffuse quickly enough to the surface, so that emission becomes insufficient. In order to achieve maximum life, the heater voltage is required to be within +5% and -10% of the limits

prescribed in the performance specifications. (See Appendix 1.)

Anode Supply

A magnetron is generally represented by an equivalent circuit for purposes of anode supply design. The input of a magnetron is characterized by its Hartree voltage (V_{th}) and its dynamic resistance (R_m).

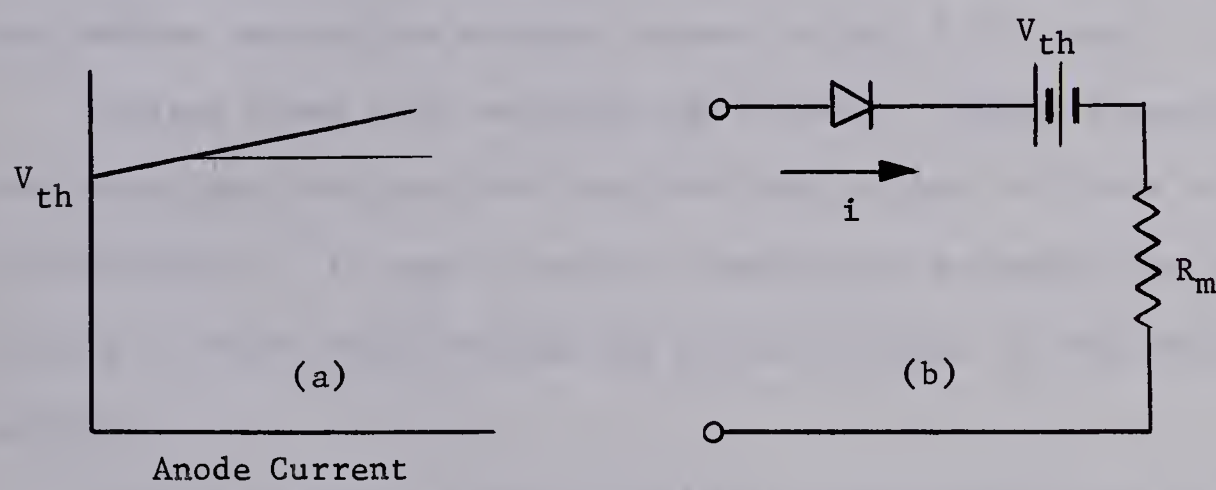


Fig. 1.2 (a) Anode Input Characteristic
(b) Magnetron Equivalent Circuit

A DC supply could be used to bias the anode. The anode current would then be controlled by using either a variable DC voltage supply or a fixed voltage supply in series with a variable resistor.

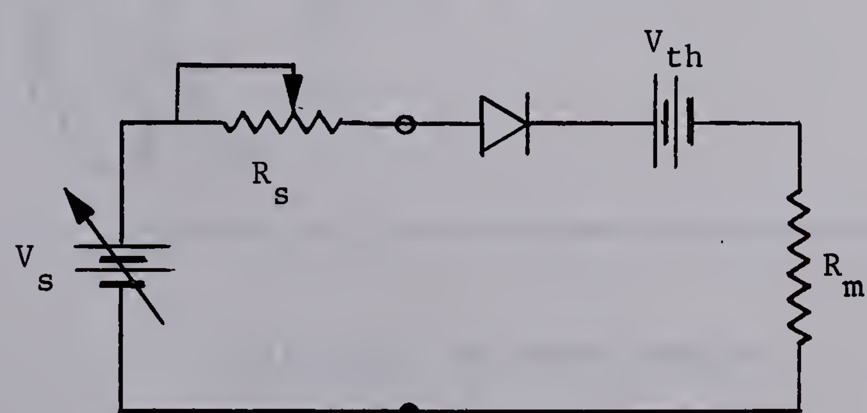


Fig. 1.3 A DC Anode Supply

The use of a pure DC supply is not recommended, however, because it usually results in lower magnetron efficiency. Magnetrons operate more efficiently at high anode current. Therefore, it is desirable to operate the magnetron at high current during part of a voltage cycle, instead of continuously at a lower current. The Philips 7292 magnetron, for example, can be operated at instantaneous anode currents as high as 2.0 amps but the maximum permissible average current is only 0.750 amps.

Various types of AC supplies can be used. Single phase, two phase, and three phase systems have been used with either full wave or half wave rectification. In some locations, regulations governing the unbalanced loading of three phase systems may dictate the use of only three phase systems.

With an AC supply either a resistor or an inductor can be used to limit and control the anode current. There are advantages to either method. W. Schmitt[8] has analyzed the use of resistive current limiting in a single phase anode supply. A summary of his findings follows.

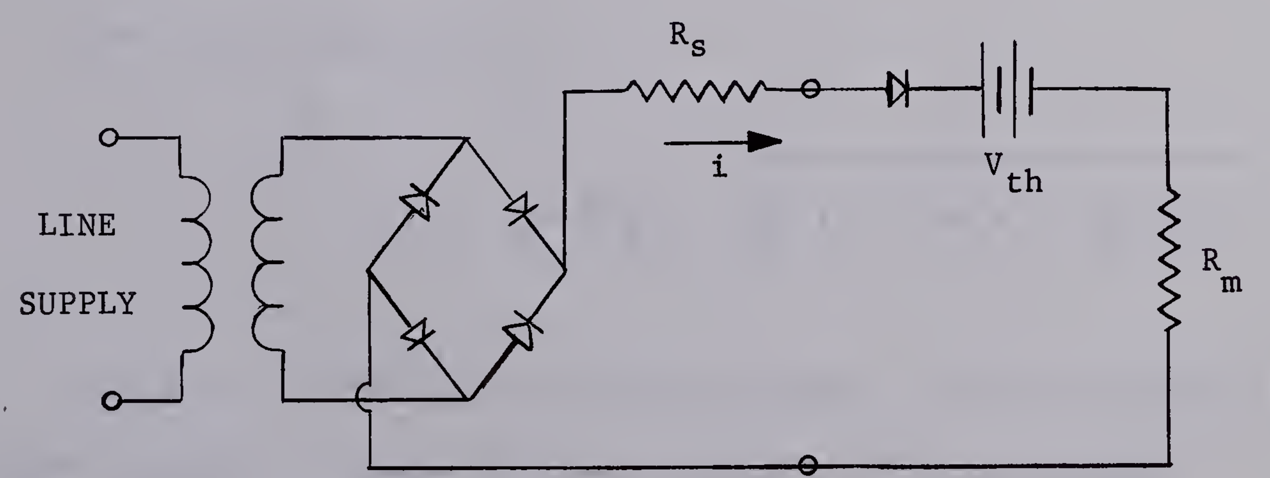


Fig. 1.4 AC Anode Supply

The half conduction angle θ can be defined in terms of the Hartree voltage (V_{th}) of the magnetron and the peak input voltage ($\hat{v}_o$)

$$\theta = \cos^{-1} \frac{V_{th}}{\hat{v}_o} \qquad \ldots 1.1$$

The instantaneous current is

$$I = \frac{V_{th}}{R_m + R_s} \cdot \frac{\cos(wt) - \cos(\theta)}{\cos(\theta)} \qquad \ldots 1.2$$

for $-\theta < wt < \theta$.

Using Eq. 1.2, the peak, average, and r.m.s. currents can be calculated. The peak current is

$$i_p = \frac{V_{th}}{R_m + R_s} \cdot \frac{1 - \cos(\theta)}{\cos(\theta)} \qquad \ldots 1.3$$

The average current is

$$i_A = \frac{V_{th}}{R_m + R_s} \cdot \frac{\sin\theta - \theta\cos\theta}{\pi\cos\theta} \qquad \ldots 1.4$$

The r.m.s. current is

$$i_{rms} = \frac{V_{th}}{R_m + R_s} \cdot \sqrt{\frac{\theta}{2\pi}(3 + \tan^2\theta) - \frac{3}{2\pi}\tan\theta} \qquad \ldots 1.5$$

Using the expression for average current, the sensitivity of average anode current to changes in input voltage can be calculated.

$$\frac{di_A}{i_A} = \frac{\sin(\theta)}{1 - \theta\cot(\theta)} \frac{d\hat{v}_o}{\hat{v}_o} \qquad \ldots 1.6$$

A similar analysis has been carried out for the case where an inductor is used for current limiting.[9] The following table provides a direct comparison between resistive current limiting and inductive current limiting.

TABLE 1.1

Comparison of Resistive and Inductive Current Limiting

1. Resistive Current Limiting

Advantages

(a) A resistor is light in weight and inexpensive.

(b) Variable resistors can be used.

Disadvantages

(a) A resistor dissipates power and is therefore less efficient

(b) High transformer voltages are necessary for large conduction angles.

2. Inductive Current Limiting

Advantages

(a) Increased power supply efficiency.

(b) Increased conduction angles at lower transformer voltages.

(c) Better crest factor of the current waveshape which permits the use of thinner transformer wire.

(d) Reduced peak current.

Disadvantages

(a) Weight and cost of the inductor.

(b) Difficulty in controlling anode current.

(c) Sensitivity of the anode current to changes in line voltage.

A saturable core reactor is generally used with both resistive and inductive current limited supplies to provide regulation to changes in line voltage and to control the anode current. A saturable core reactor supply will operate with a regulation coefficient of 0.5 (i.e. a 5% change of anode current for a 10% change of line voltage).

1.6 An S.C.R. Controlled Magnetron Power Supply

S.C.R.'s are used to control the conduction angle of current in an AC system. They can be fired with low voltage transistorized circuitry. With appropriate circuitry, the firing angle can be controlled with a DC input voltage. Hence, feedback systems can be used to provide good regulation to line voltage changes.

Saturable core reactor controlled supplies have many undesirable features. An S.C.R. controlled supply could offer the following improvements.

(a) A reduction in size and weight of the supply.

(b) Better regulation to changes in line voltage by means of feedback systems.

(c) Faster response to control signals.

(d) Feedback regulation of r.f. field strengths.

(e) Voltage control, thereby permitting programmed operation.

(f) Protection of individual tubes in a multiple tube system.

(g) More precise control of the heater temperature.

1.7 Object of the Work

The advantages of using solid state control systems in the power supply for a continuous wave magnetron are investigated in this thesis. A supply is designed and constructed for a Philips 7292, two kilowatt, continuous wave magnetron. The supply is tested and the performance compared to that of a saturable core reactor controlled supply.

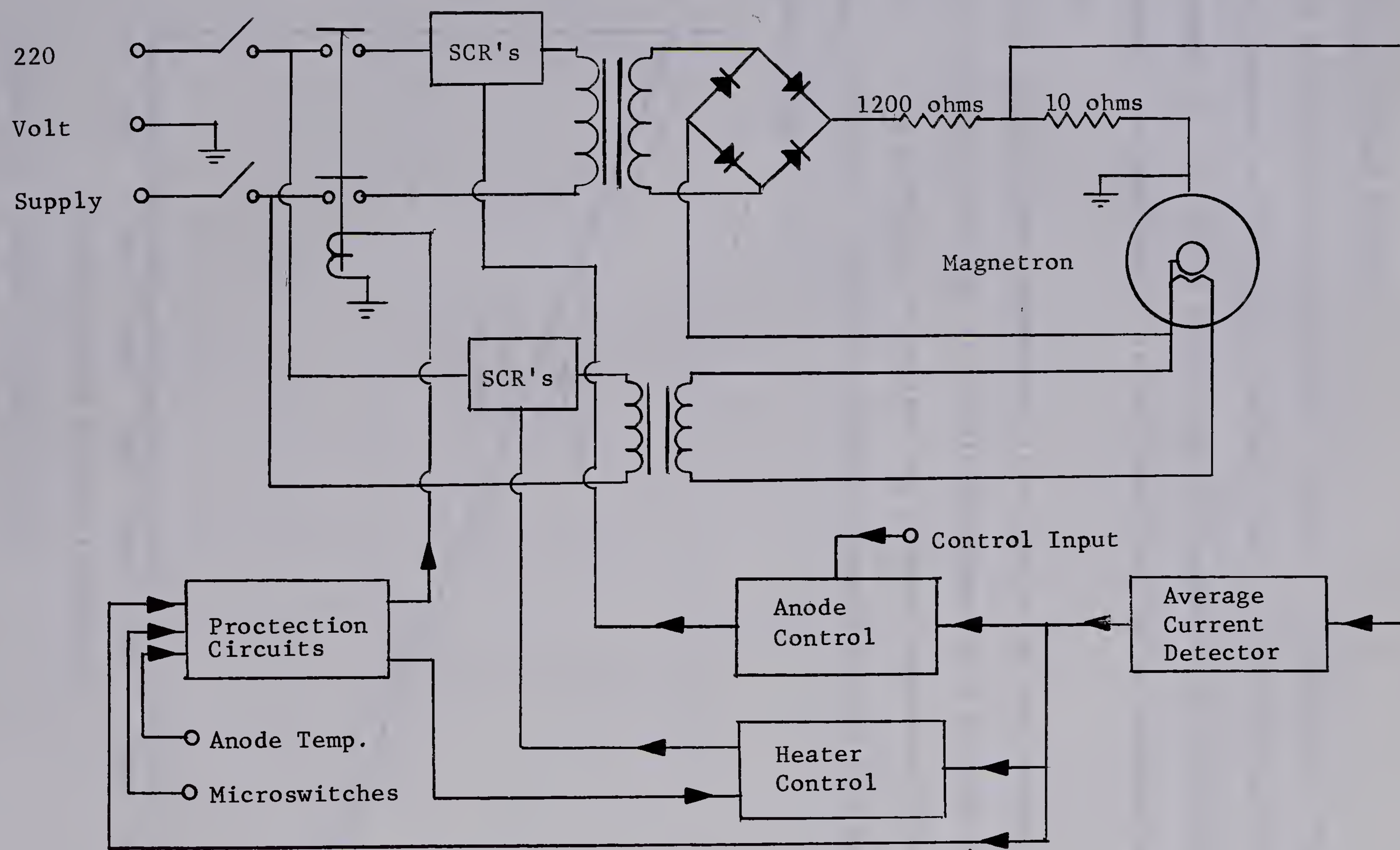


Fig. 1.5. Block Diagram of the Solid State Controlled Power Supply

2. HEATER CONTROL

2.1 Heater Voltage Schedule

The purpose of the heater control system is to maintain a constant cathode temperature throughout the operating range of the magnetron. Back bombardment of the cathode occurs when the magnetron is oscillating and it is necessary to compensate this additional heating by reducing the heater voltage.

Since electron back bombardment is a function of the average anode current, the heater voltage is usually programmed as a function of anode current. It is difficult to monitor the cathode temperature directly for purposes of control. The heater voltage schedule for the 7292 magnetron is shown in Fig. 2.1.

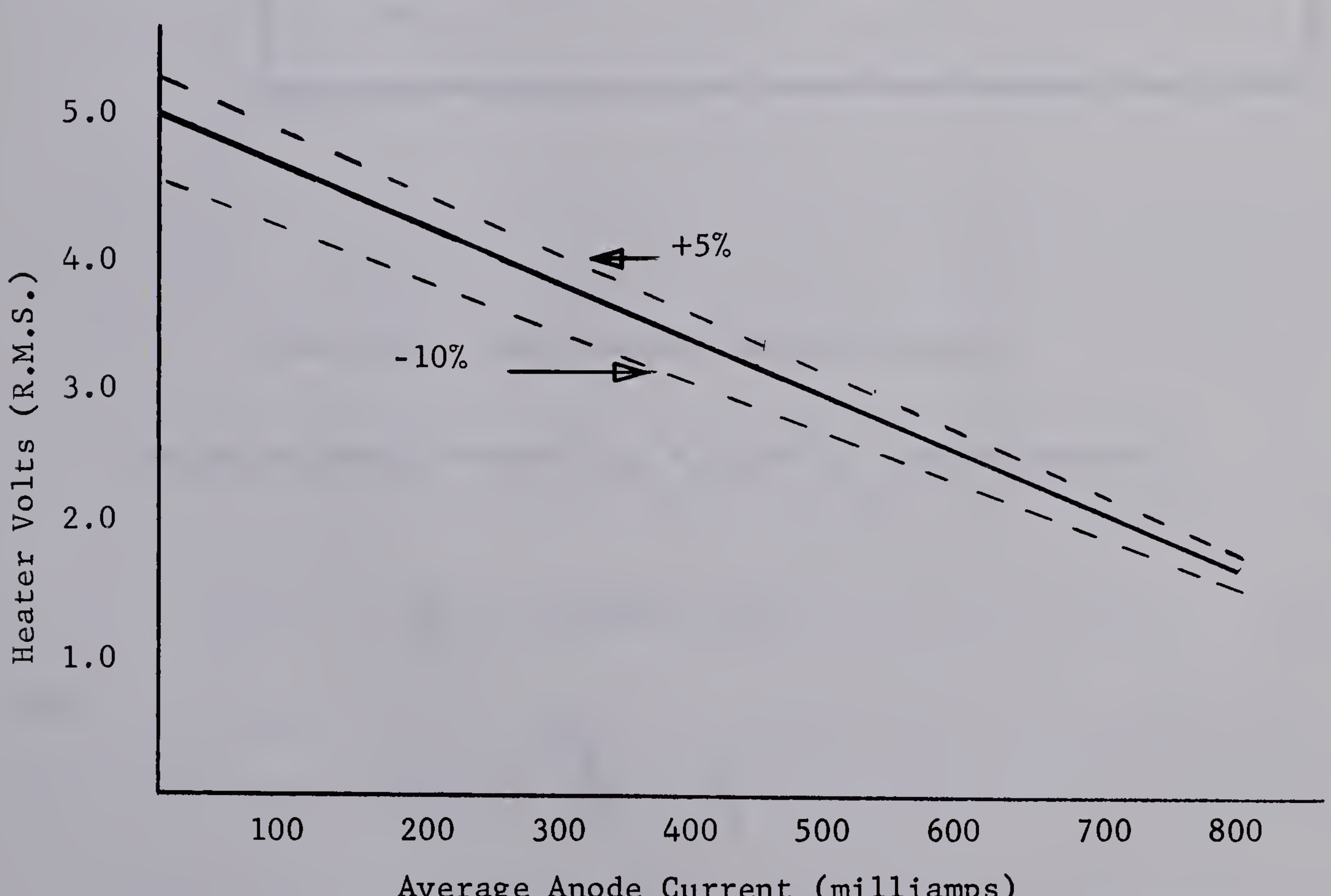


Figure 2.1. Heater Voltage Schedule.

2.2 Heater Control System

An average current detector circuit is used to produce a DC voltage, which is proportional to the average value of the anode current. This DC voltage is used in the heater control system to back off the heater voltage as the anode current is increased. The basic idea behind the system is illustrated in Fig. 2.2.

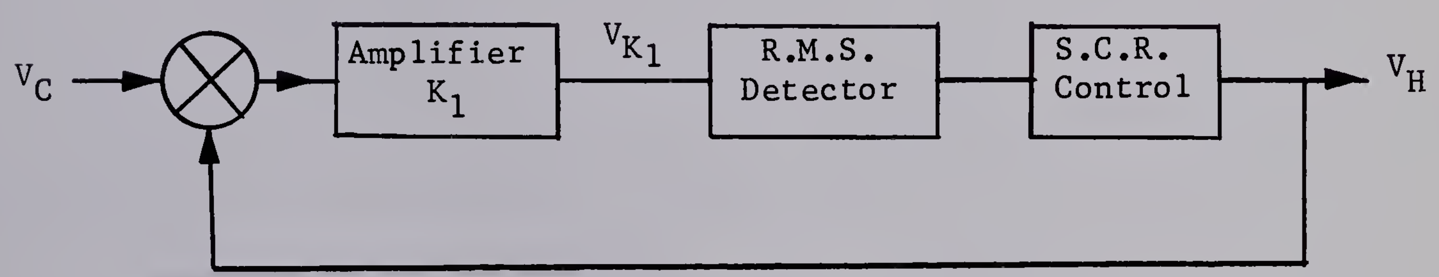


Fig. 2.2 Basic Heater Control System

The relationship between V_H, V_C, and V_{K_1} can be derived.

$$V_{K_1} = K_1(V_H + V_C) \qquad \text{... 2.1}$$

and

$$V_H = \frac{V_{K_1}}{K_1} - V_C \qquad \text{... 2.2}$$

The system is designed to regulate the voltage V_{K_1} at a selected value. If $V_{K_1} = K_1$ (5.0 volts) and the correct proportionality constant relating V_C and the average value of anode current is chosen, then Eq. 2.2 will yield

the desired relationship shown in Fig. 2.1.

One error exists in the system as it stands, however. The sum of the individual r.m.s. values of V_H and V_C is not necessarily equal to the r.m.s. value of the combined waveshape $(V_H + V_C)$. A mathematical analysis is performed (see Appendix 5) to set conditions for the r.m.s. addition of waves. It is shown that the r.m.s. addition of waves can only be performed if the waves are identical in shape (not necessarily magnitude) and are in phase.

The system shown in Fig. 2.3 provides the desired relationship between V_C and V_H and the conditions for r.m.s. addition of waves are satisfied.

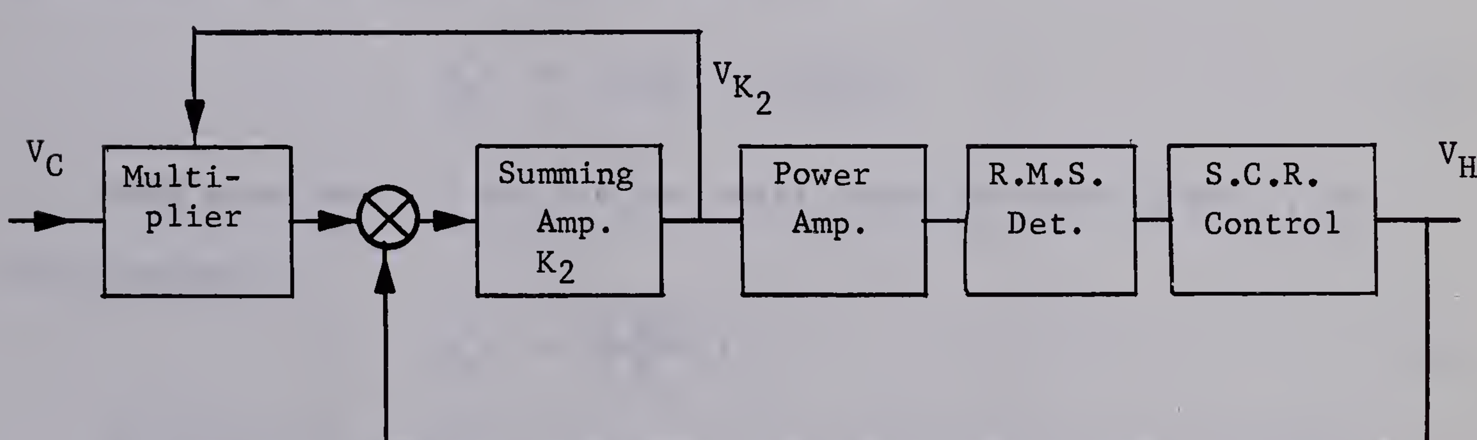


Fig. 2.3 Block Diagram of the Heater Control System

The system acts to regulate the voltage V_{K_2} to a selected value, in this case

$$V_{K_2} = K_2 \text{ (5.0 volts)} . \qquad \text{... 2.3}$$

The relationship between V_H and V_C can be calculated for the system shown in Fig. 2.3.

$$V_{K_2} = K_2(V_M - V_H) \qquad \ldots\ 2.4$$

But

$$V_M = V_C V_{K_2} \qquad \ldots\ 2.5$$

Therefore

$$V_H = \frac{V_{K_2}}{K_2}(1 - K_2 V_C) \qquad \ldots\ 2.6$$

Since

$$\frac{V_{K_2}}{K_2} = 5.0 \text{ volts.} \qquad \ldots\ 2.7$$

The mathematical expression for the relationship between heater voltage and average anode current shown in Fig. 2.1 is

$$V_H = 5.0(1 - 0.825 I_A) \qquad \ldots\ 2.8$$

From equations 2.7 and 2.8 the relationship between V_C and I_A can be calculated.

$$V_C = \frac{0.825}{K_2} I_A \,. \qquad \ldots\ 2.9$$

In the actual system the output of the average current detector, $V_{A.C.}$, is related to the average anode current, I_A, by

$$V_{A.C.} = 10\, I_A \,, \qquad \ldots\ 2.10$$

and adjustments are made in the multiplier to set the correct relationship between V_C and the anode current.

Although positive feedback is used in the system, the system is stable because the maximum value of open loop gain is 0.65.

The individual circuits shown in Fig. 2.3 are discussed in the following sections. Linearity of the system is discussed in the last section of this chapter.

2.3 The Summing Amplifier

A simple three transistor operational amplifier modified by the addition of an emitter follower stage and a current source, is used for the summing amplifier. The voltage gain is set at 3.7 for the signal from the multiplier. The input resistor for the heater voltage signal is made adjustable so that the heater voltage at standby (no anode current) can be adjusted to the desired value.

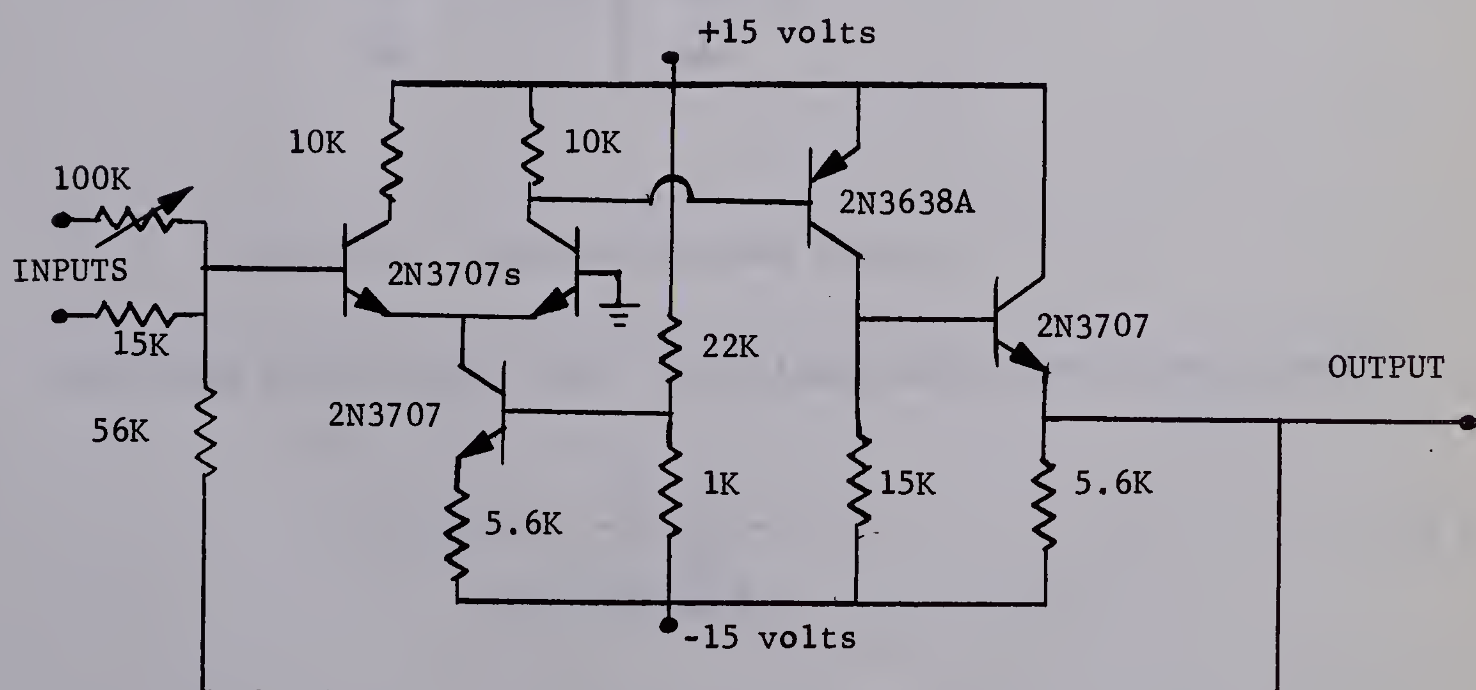


Fig. 2.4 Summing Amplifier for Heater Control System

2.4 The Multiplier Circuit

The multiplier circuit is required to multiply a 60 Hz wave and harmonics by a DC voltage. Phase shift requirements are important and the lower corner frequency is therefore set far below 60 Hz. Transients in the output due to changes in the DC input will affect the transient response of the system and an effort is made to reduce them.

A simple circuit was originally considered for the multiplier, but was later rejected in favour of a more sophisticated design based on the same principles. The first circuit is discussed before the final design is considered.

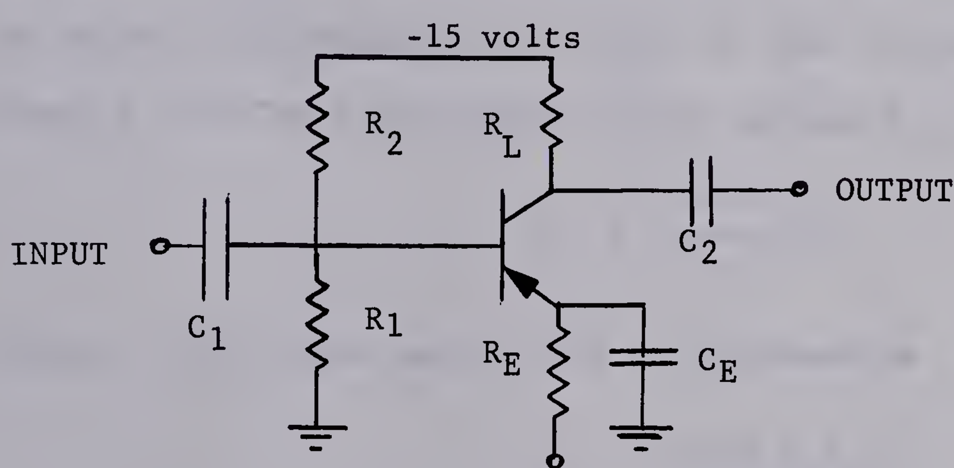


Fig. 2.5 First Multiplier Circuit

Neglecting second order terms, the voltage gain of the above circuit is

$$A_v = \frac{-R_L}{h_{ib} + \dfrac{R_E}{1 + sR_EC_E}} \qquad \ldots 2.11$$

If C_E and R_E are made large so that $h_{ib} << R_E$ and $1/wC_E << h_{ib}$, Eq. 2.12 can be expressed as

$$A_v = \frac{-R_L}{h_{ib}} \qquad \ldots\ 2.12$$

If the collector current (I_C) is small the following approximation for h_{ib} is valid. I_C is expressed in milliamps.

$$h_{ib} = \frac{26}{I_C} \qquad \ldots\ 2.13$$

Substituting Eq. 2.13 into Eq. 2.12 gives

$$A_v = \frac{-R_1 I_C}{26} . \qquad \ldots\ 2.14$$

The voltage gain is then proportional to the DC collector current.

Now if the bias resistors are chosen so that the emitter voltage is zero, the current I_C will be proportional to the voltage V_{DC} (volts)

$$I_C = \frac{1000\ V_{DC}}{R_E} \qquad \ldots\ 2.15$$

Finally, the voltage gain A_v can be expressed as

$$A_v = \frac{-1000\ R_L V_{DC}}{26\ R_E} \qquad \ldots\ 2.16$$

The first multiplier circuit has three faults, however.

(a) C_E must be a very large capacitor operating at essentially zero volts.

(b) Any change in V_{DC} will cause a transient in the output. Because of the phase shift requirements of the multiplier, the coupling capacitors would be large and transients would have long time constants.

(c) Any change in ambient temperature will cause a -2mv/°C change in V_{BE} and therefore in V_E.

In order to temperature compensate the circuit and to eliminate capacitor C_E, the following circuit is used.

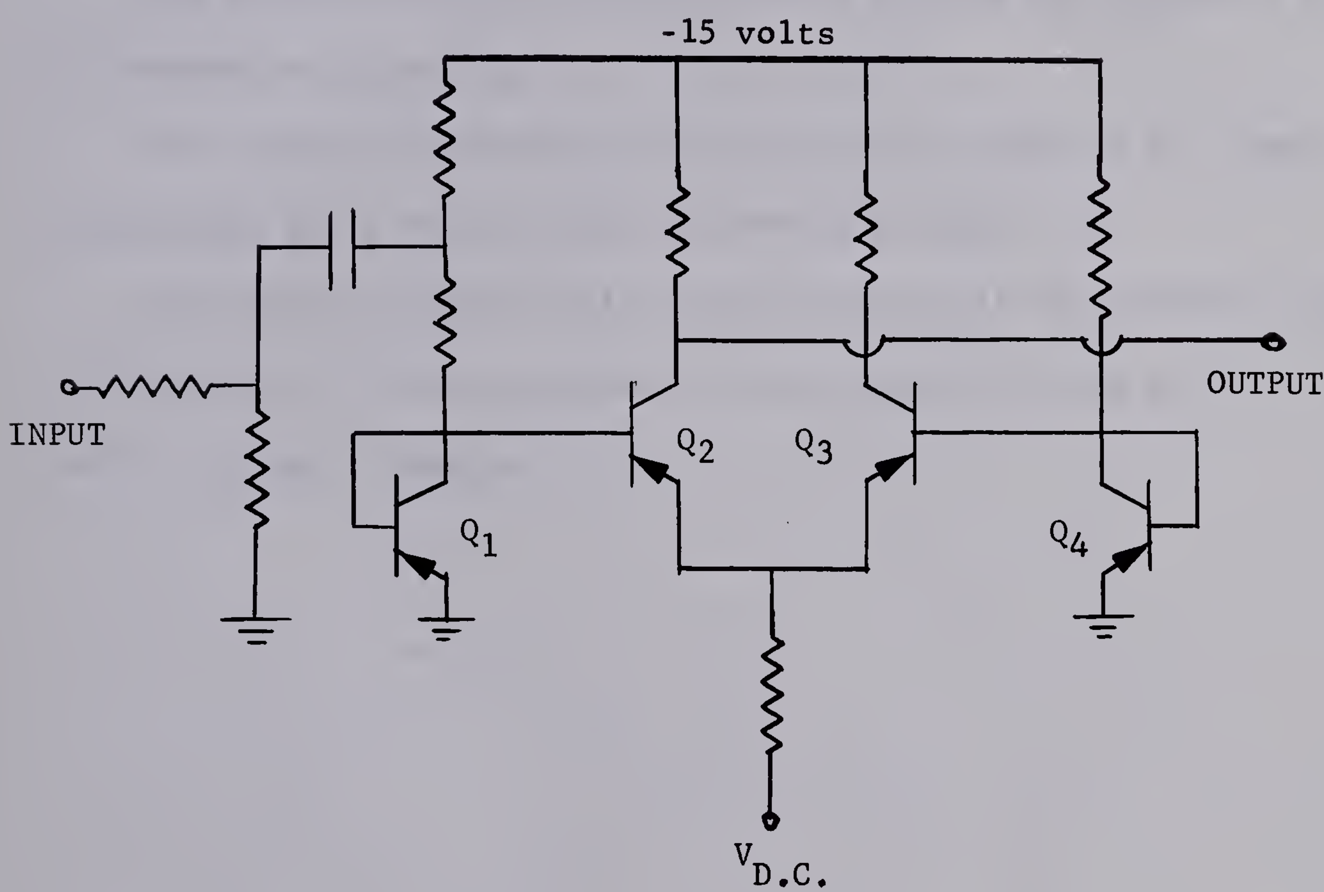


Fig. 2.6 First Differential Stage of Multiplier

This circuit compensates the thermal drift of transistors Q_2 and Q_3 by the addition of transistors Q_1 and Q_4. The capacitor C_E of the previous circuit is eliminated by the use of the differential stage.

Another advantage of the above circuit is that the output signal will occur as a differential signal in a double ended output, but changes in V_{DC} will appear as a common mode signal in the same output. A second

differential stage with good common mode rejection is therefore added to eliminate any transients resulting from changes in V_{DC}.

Resistor R_E is made variable so that a final adjustment could be made to the heater control system once the power supply is operating.

The transistors used are selected to balance the collector currents in transistors Q_2 and Q_3.

The effect of distortion is considered in Appendix 2. The circuit is designed for a maximum peak distortion of about 5%.

The effect of phase shift on the accuracy of the system is considered in Appendix 3. Coupling capacitors are chosen to give an overall phase shift of about 2 degrees.

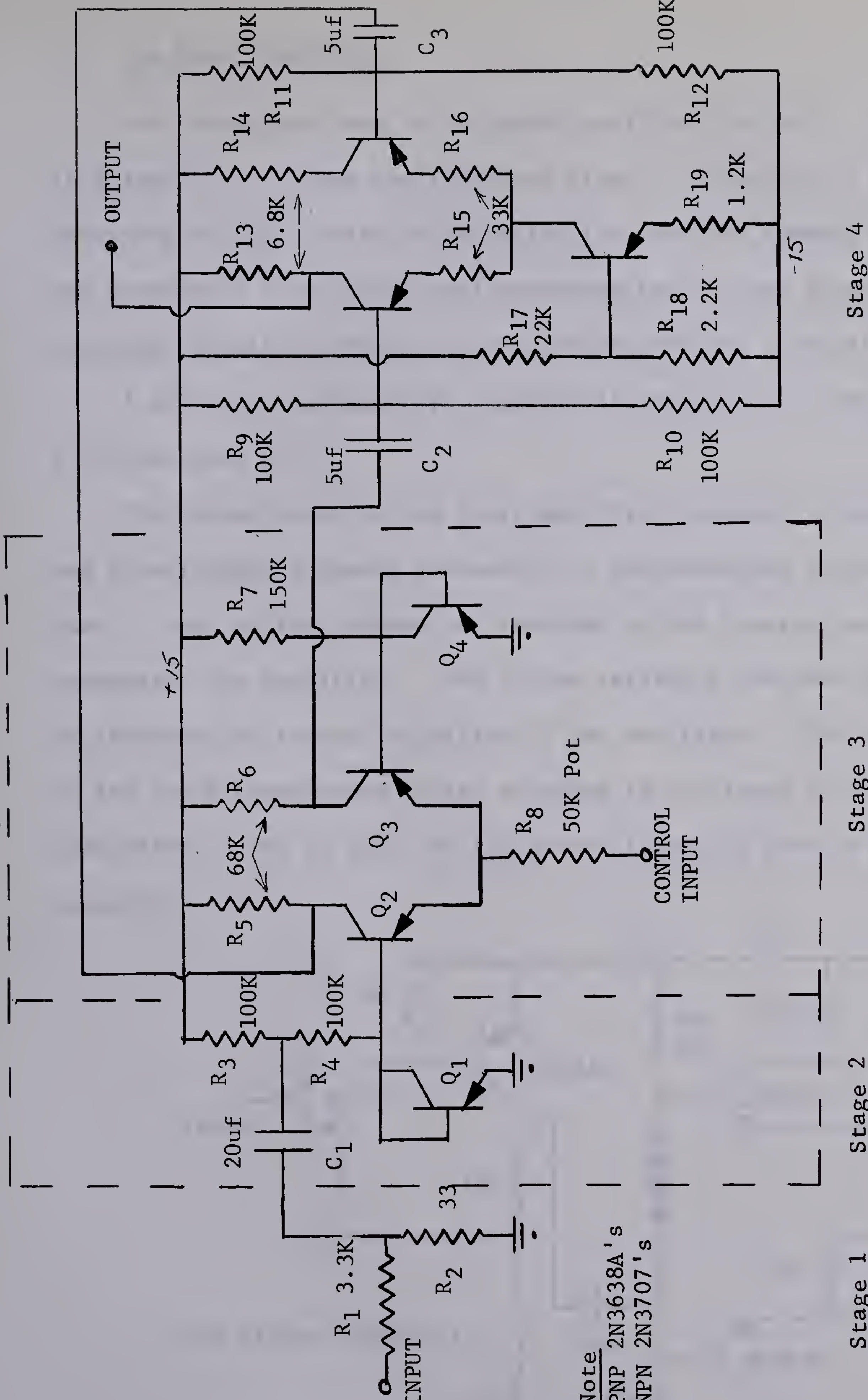


Fig. 2.7. The Multiplier Circuit.

2.5 The Power Amplifier

Two stages are used in the power amplifier circuit. The first stage is primarily a voltage amplification stage. It permits a reduction of the operating voltage levels of the multiplier and the summing amplifier circuits, and provides a DC voltage level necessary for the use of an electrolytic coupling capacitor between its own output, and the input of the second stage.

A modified complementary compound is used as the first stage. It has a voltage gain of 1.3.

The second stage of the power amplifier contains a complementary compound and a Darlington compound arranged in a complementary emitter follower configuration. Four silicon diodes are included in the biasing network to thermally compensate the amplifier. Two 10 ohm resistors are used as emitter resistors to increase the thermal stability of the amplifier. The quiescent DC current of the power transistors is set at about 15 milliamps to reduce cross over distortion. The DC level of the output is set to zero by adjusting the biasing resistors.

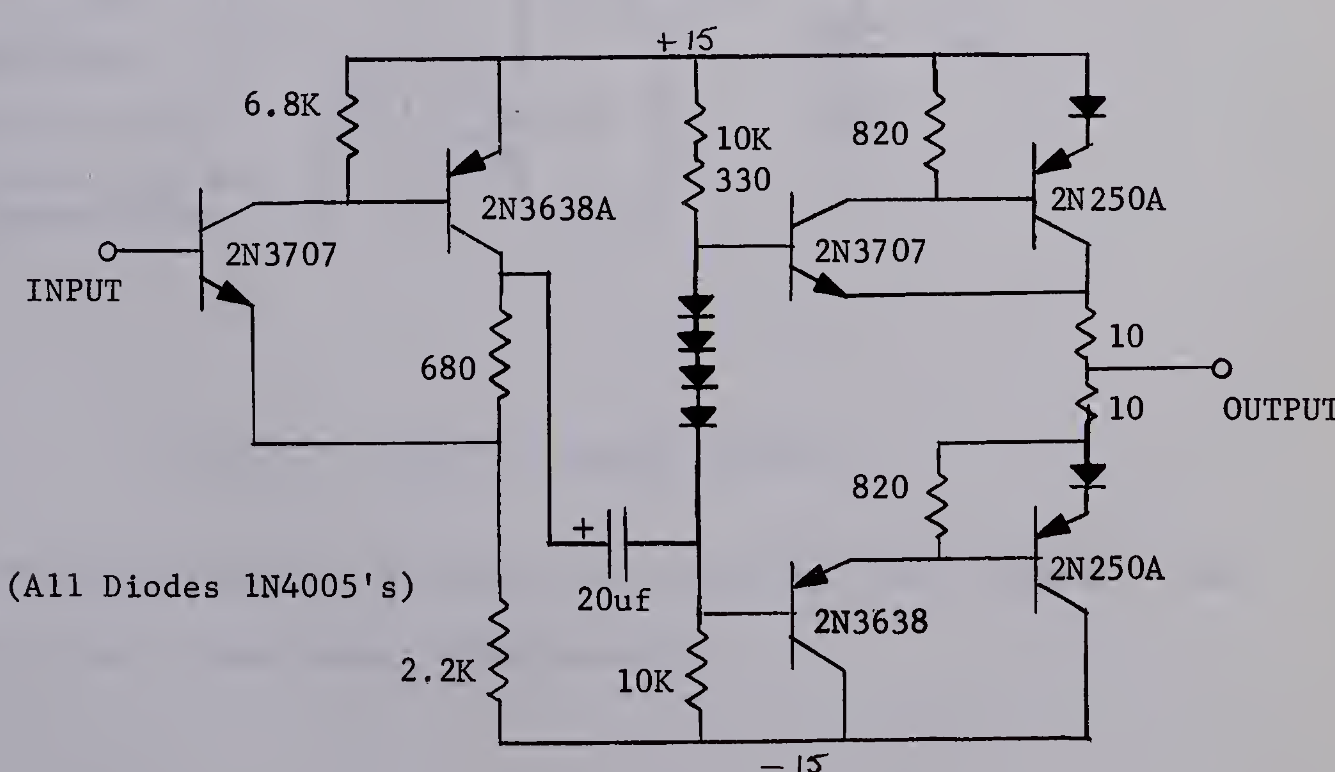


Fig. 2.8 Power Amplifier

2.6 The R.M.S. Detector

A tungsten filament light bulb and phototransistor are used as an r.m.s. detector. The light intensity varies as the power supplied to the bulb and therefore as the r.m.s. voltage applied.

The bulb is operated at about one-half its rated voltage. Operating a tungsten bulb at one half its rated voltage will extend its life by a factor of 10,000.[10] The bulb used is rated at 6.3 volts and 0.25 amps.

The light detector element is a Texas Instrument LS-400 NPN phototransistor. This device is arranged in a circuit which provides a DC output voltage which decreases as the light intensity increases.

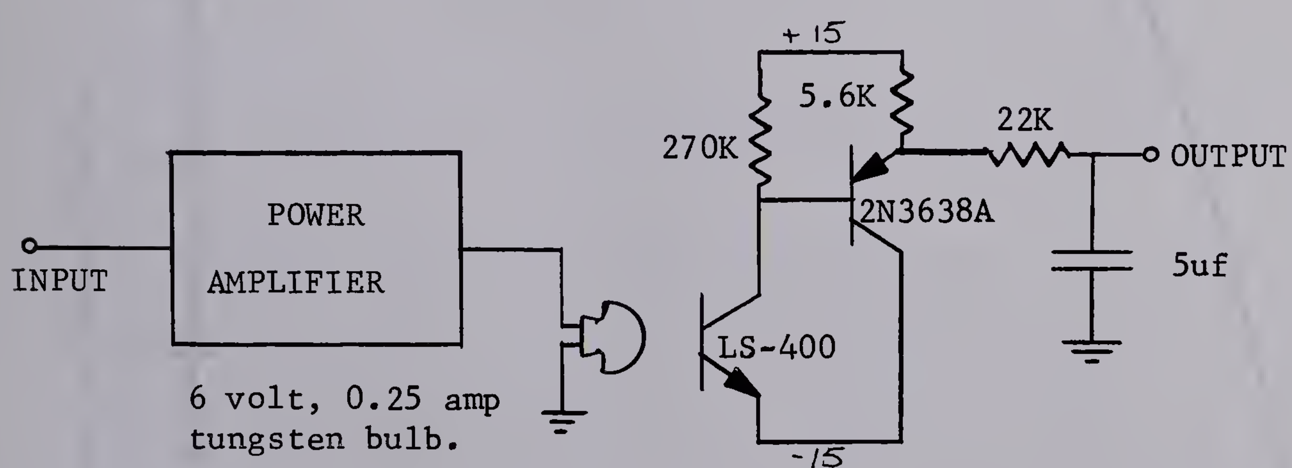


Fig. 2.9 R.M.S. Detector Circuit

An RC filter network is used at the output to filter ripple. The values of R and C were chosen experimentally.

The light bulb and phototransistor are mounted about one half inch apart. They are placed in a closed metal box to isolate the system from ambient light.

The transfer function of the r.m.s. detector circuit has been obtained experimentally. It is shown below.

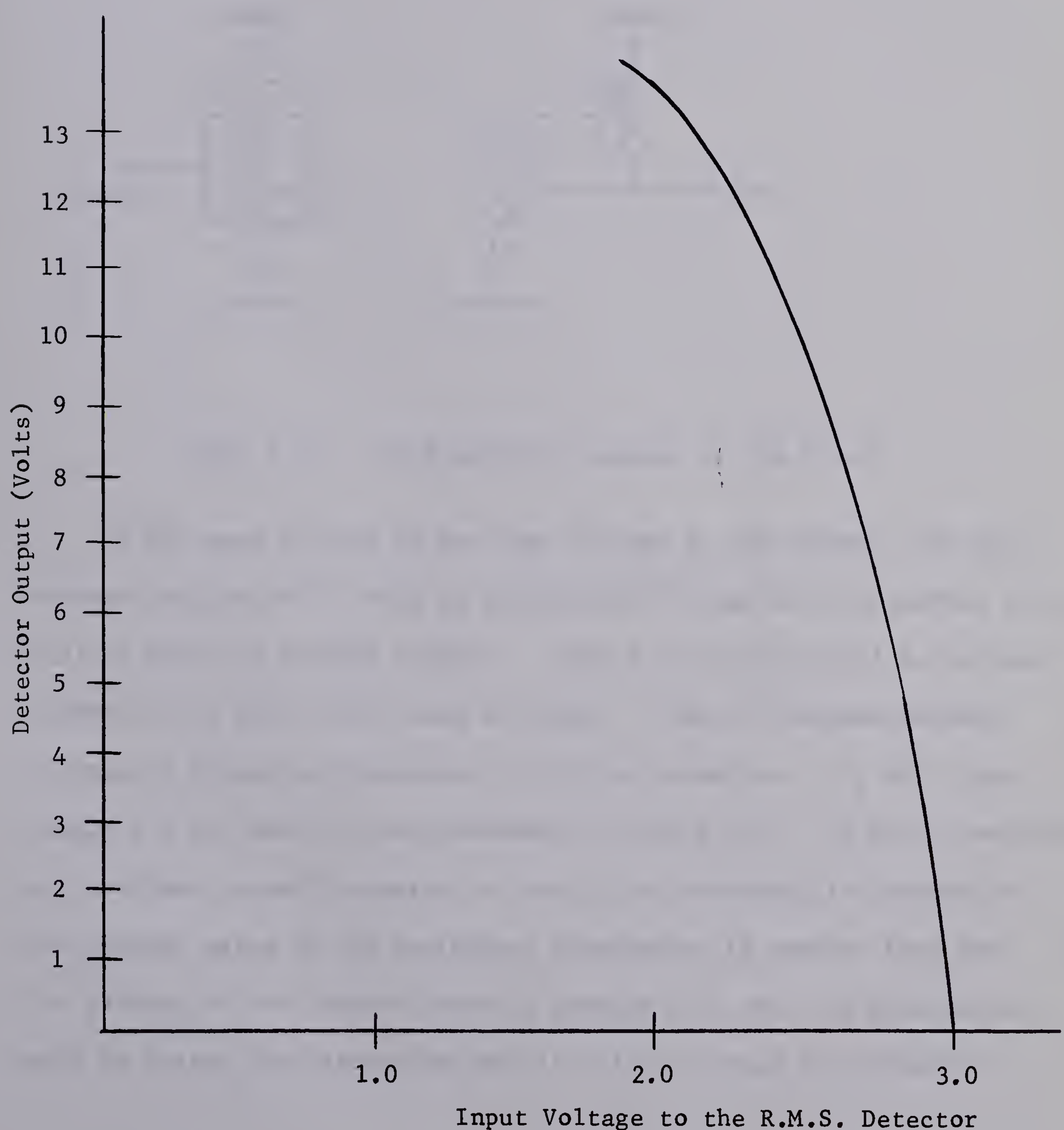


Fig. 2.10. Transfer Function of the R.M.S. Detector.

2.7 The S.C.R. Trigger Circuit

An S.C.R. consists of four layers of alternate N-type and P-type semiconductor materials. For descriptive purposes the four-layer structure is generally split into two three-layer structures. (Two transistor analogy.)[11]

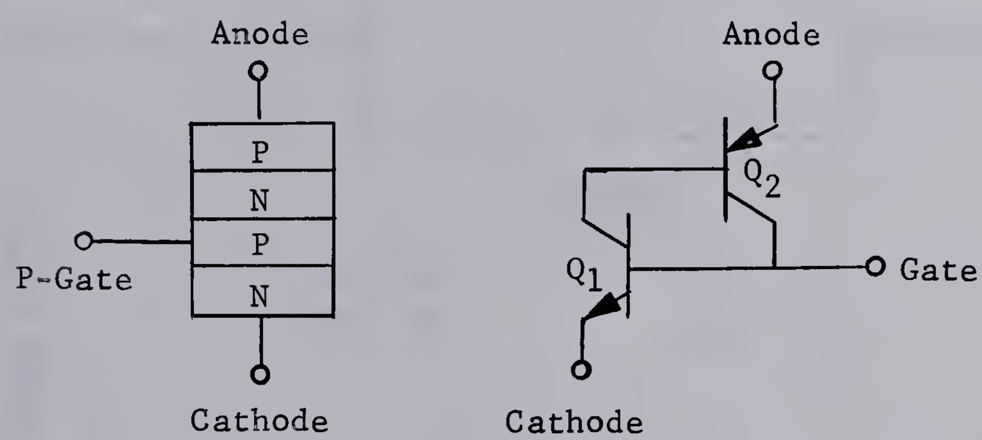


Fig. 2.11 Two Transistor Analogy of the S.C.R.

If the gate is held at the same voltage as the cathode, the base emitter junction of Q_1 will be in the cutoff stage and the current in Q_1 will be only the leakage current. With Q_1 off, there will be no base current in Q_2 and it will also be cutoff. Now if the gate-cathode voltage is raised sufficiently, Q_1 will be turned on. Q_1 will then supply all the base current necessary to turn Q_2 on. Q_2 will then supply all the base current necessary to keep Q_1 on providing the product of the current gains of the individual transistors is greater than one. If the product of the current gains is greater than one, the transistors will be driven into saturation and it will no longer be necessary to

supply gate current. The device will stay in this conducting state until the current is reduced by external circuitry to the extent that the current gain

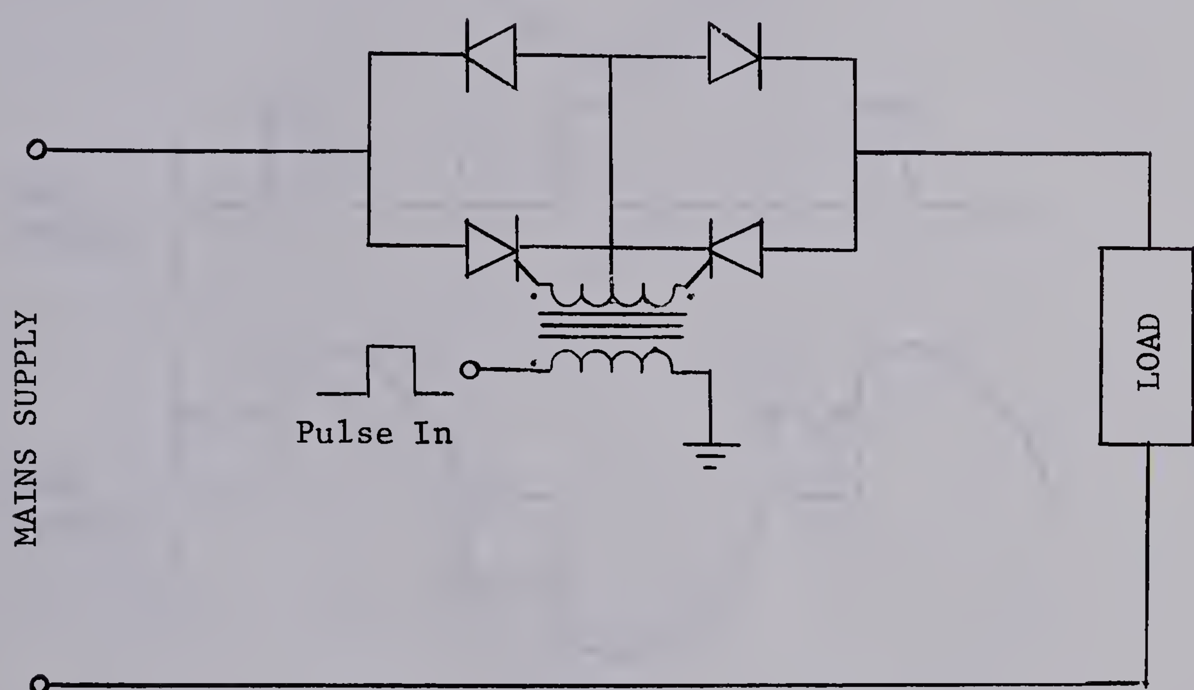


Fig. 2.12 Arrangement of S.C.R.'s

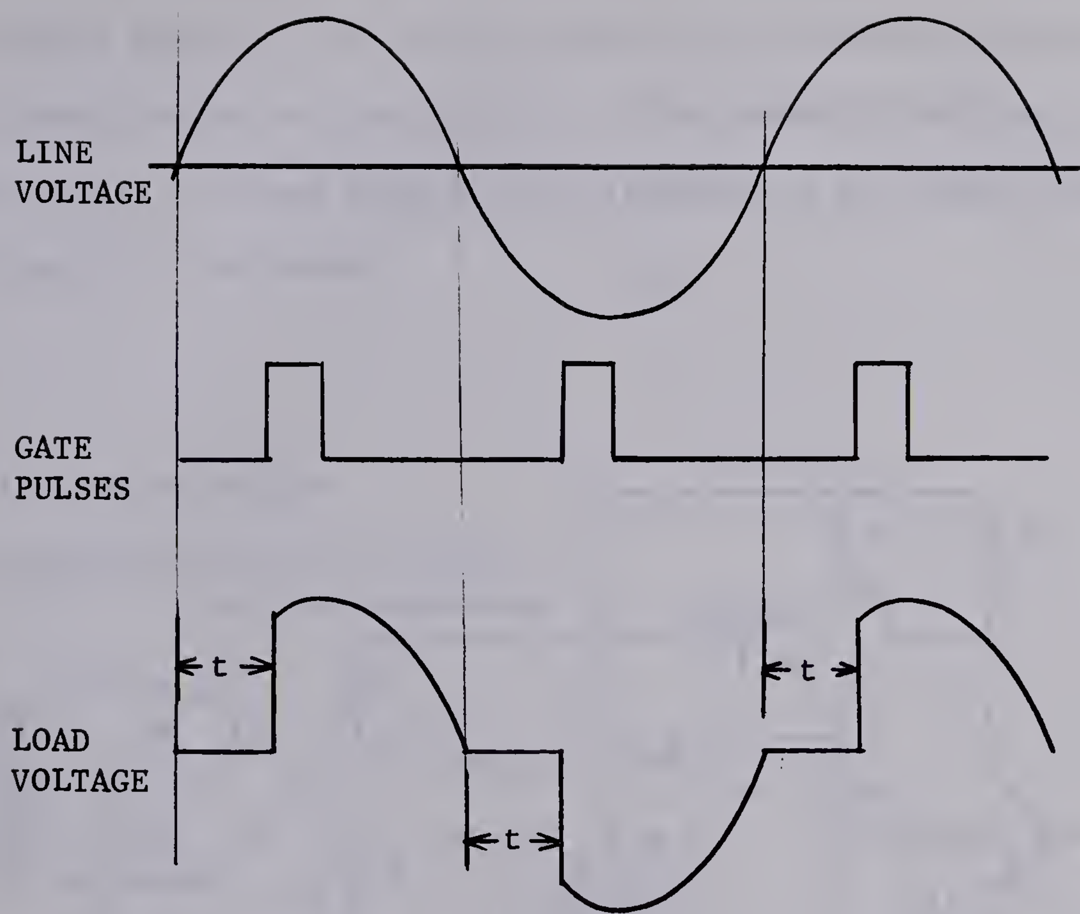


Fig. 2.13 S.C.R. Circuit Wave Shapes

In order to switch the S.C.R. into its conducting state it is only necessary to supply gate current until the device conducts with a large enough current to saturate the transistors. With resistive loads a small pulse of only a few microseconds is usually sufficient. With inductive loads, however, the build up of current in the device is much slower and longer pulses are necessary.

The S.C.R. is commonly used as an off-on switch in series with a load driven by an alternating voltage source. A circuit using S.C.R.'s and the waveshapes associated with this circuit are shown in Fig. 2.11 and Fig. 2.12.

In order to control the amount of power delivered to the load, the time delay before firing, shown as "t" in Fig. 2.12, is varied. The circuit shown in Fig. 2.13 is used in the heater control system to supply trigger pulses to the S.C.R.'s. The transfer function of the circuit is such that the load voltage will increase as the input voltage to the circuit is increased.

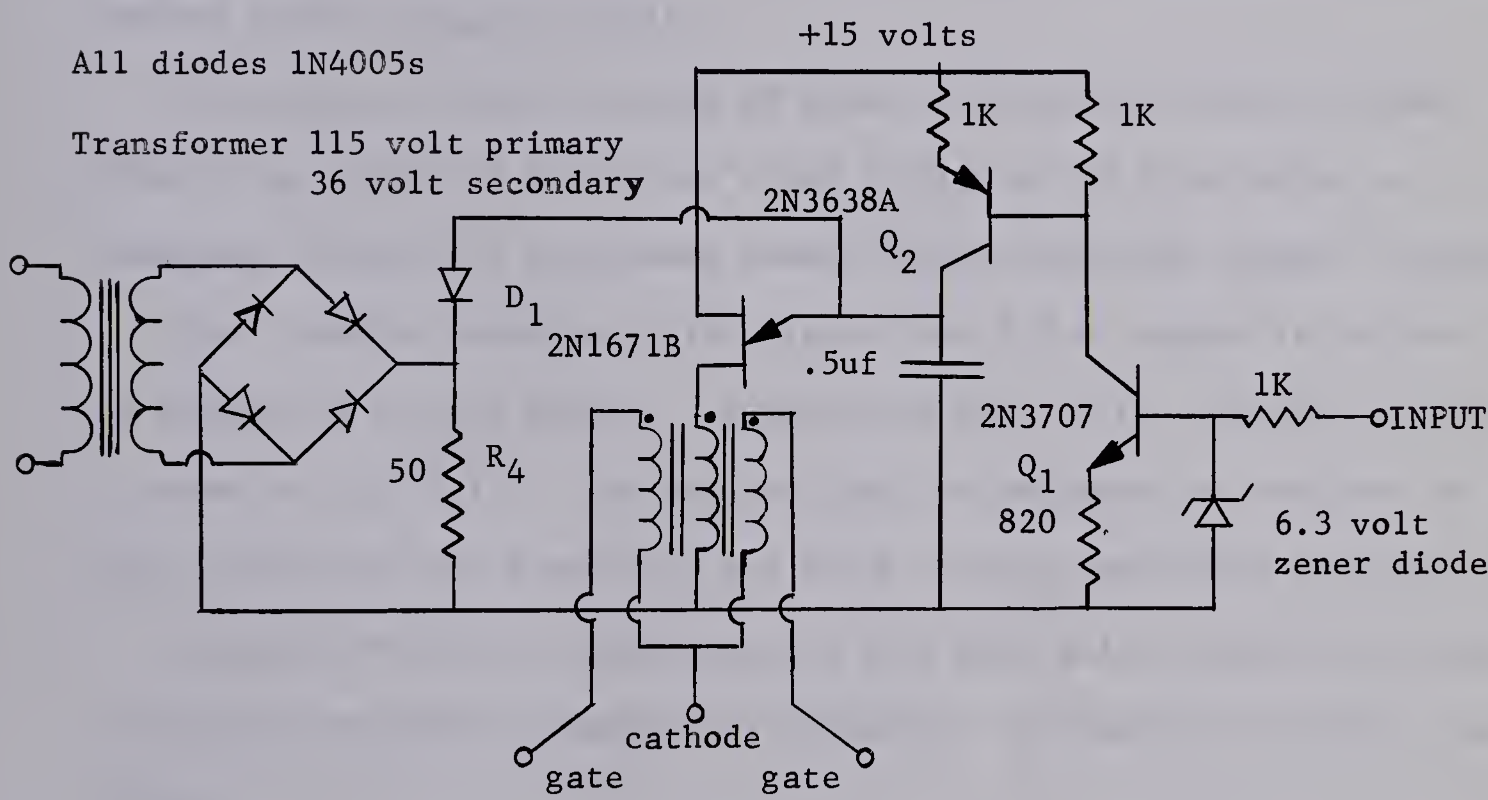


Fig. 2.14 S.C.R. Trigger Circuit

In this circuit Q_2 acts as a current source which charges up capacitor C_1. Q_1 serves as a control on the rate at which C_1 is charged. When the voltage of C_1 reaches the firing voltage V_F of the unijunction transistor, the unijunction fires delivering a pulse through the pulse transformer to the gates of the S.C.R.'s.

The circuit is made synchronous with the line voltage by discharging capacitor C_1 at the end of each half cycle of line voltage. When the voltage across resistor R_4 drops to zero, diode D_1 becomes forward biased and conducts discharging C_1 through R_4 with a time constant of 25 microseconds. The bridge rectifier and resistor can be used to synchronize more than one trigger circuit. It is also used to synchronize the anode control system trigger circuit.

The circuit delivers pulses of equal magnitude and power so that there is no danger of the S.C.R.'s not firing on the first pulse as sometimes occurs with the common zener diode unijunction trigger circuits.

The transfer function of the circuit and S.C.R. system is derived in Appendix 4 of this thesis. A graphical plot of the transfer function is shown in Fig. 2.14. The vertical axis is expressed in per cent of the voltage that would exist if the S.C.R.'s were conducting continuously.

Because the heater control system is a high gain closed loop system, no attempt was made to temperature compensate or linearize the open loop system.

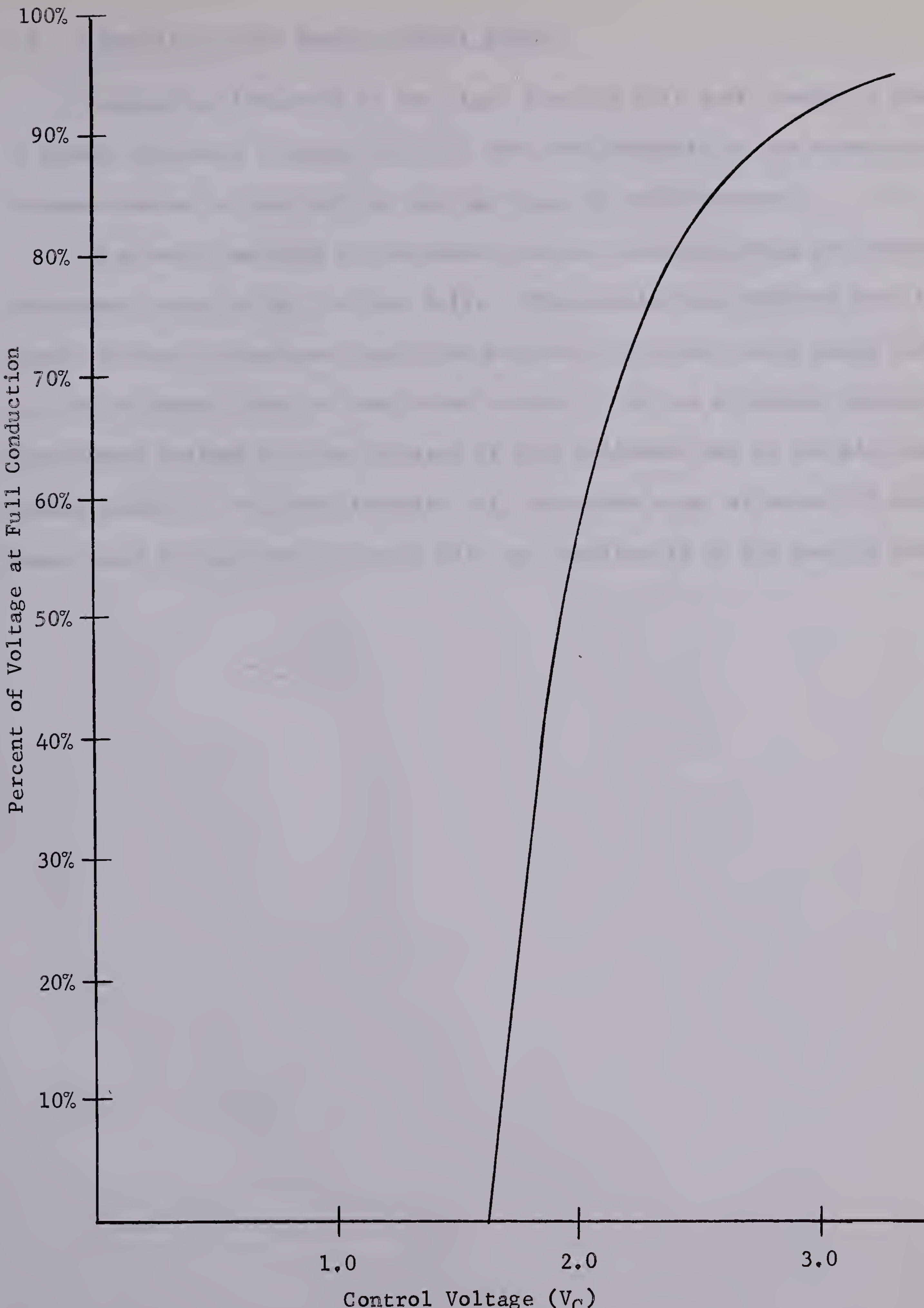
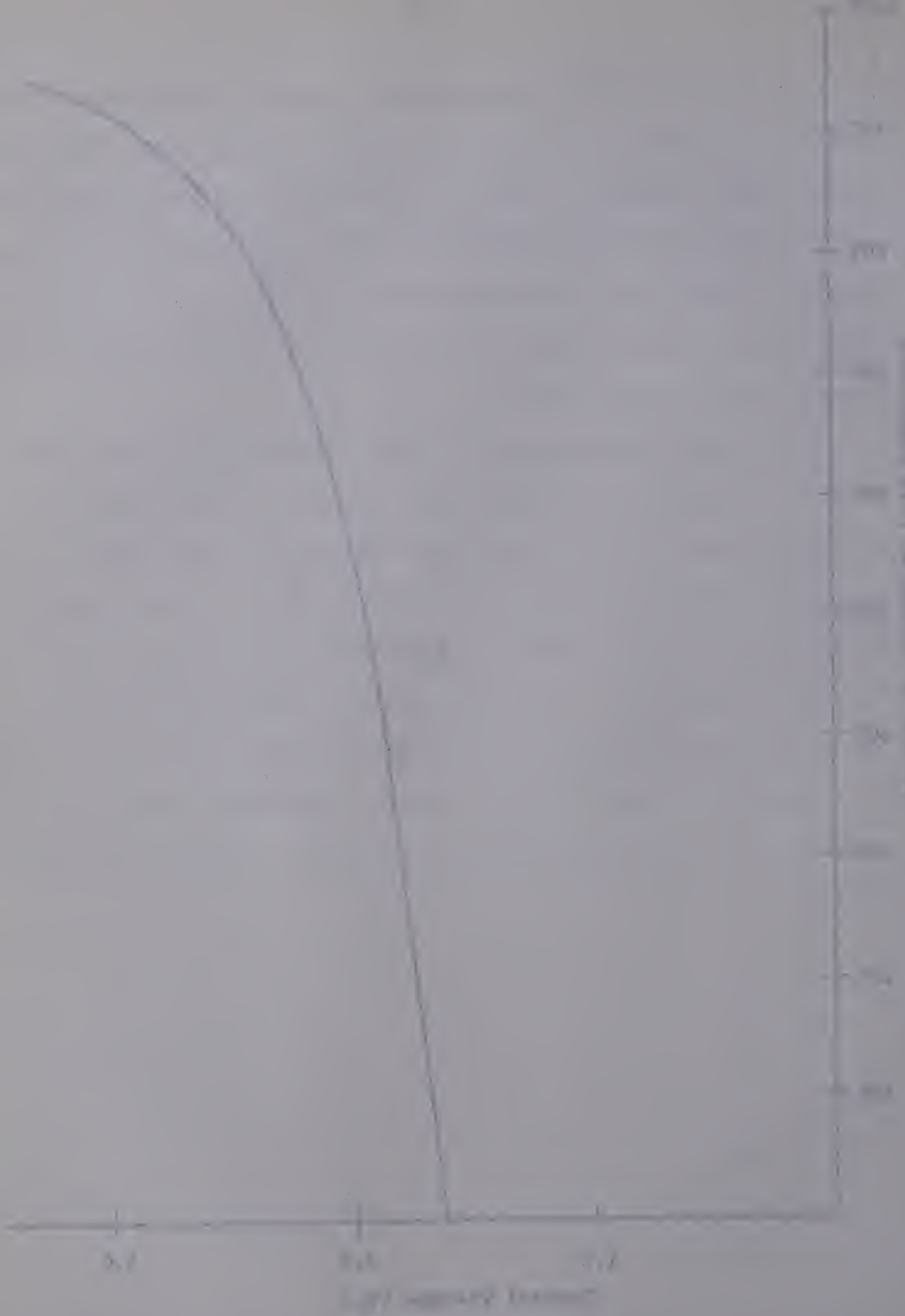


Fig. 2.15. Transfer Function of the Trigger Circuit and S.C.R.'s



2.8 Linearity of the Heater Control System

Because the intensity of the light from the bulb must change to cause a change in heater voltage, there is some non-linearity in the relationship between heater voltage and the average value of anode current.

An attempt was made to calculate this non-linearity using the transfer functions shown in Fig. 2.9 and 2.14. The calculations indicate that the input voltage to the power amplifier will be 1.7% higher at an anode current of 750 milliamps, than at zero anode current. In the alignment procedure, the heater voltage will be adjusted at zero milliamps and at 750 milliamps anode current. The non-linearity will therefore occur at about 375 milliamps anode current and the error will be less than 1% of the desired value.

3. ANODE SUPPLY

3.1 Design of the Anode Supply

A single phase, full wave rectified supply, with resistive current limiting, is used as the anode supply. Because the magnetron output power is a function of the average anode current, the supply includes a control system for regulating and controlling the average anode current. S.C.R.'s, placed in the primary circuit of the anode transformer, are used to control the average anode current.

The supply is designed so that at full power the peak and average values of anode current are 2.0 amps and 0.750 amps respectively. This choice sets the magnetron half conduction angle "Θ" from which other pertinent quantities can be calculated.[12]

Table 3.1

Peak Anode Current	2.0 amps
Average Anode Current	0.750 amps
Half Conduction Angle	52 degrees
R.M.S. Anode Current	1.1 amps
Peak Secondary Transformer Voltage	7.1 kilovolts
Current Limiting Resistor	1200 ohms
Transformer Turns Ratio	22.8
Primary Transformer Current	25.1 amps (rms)

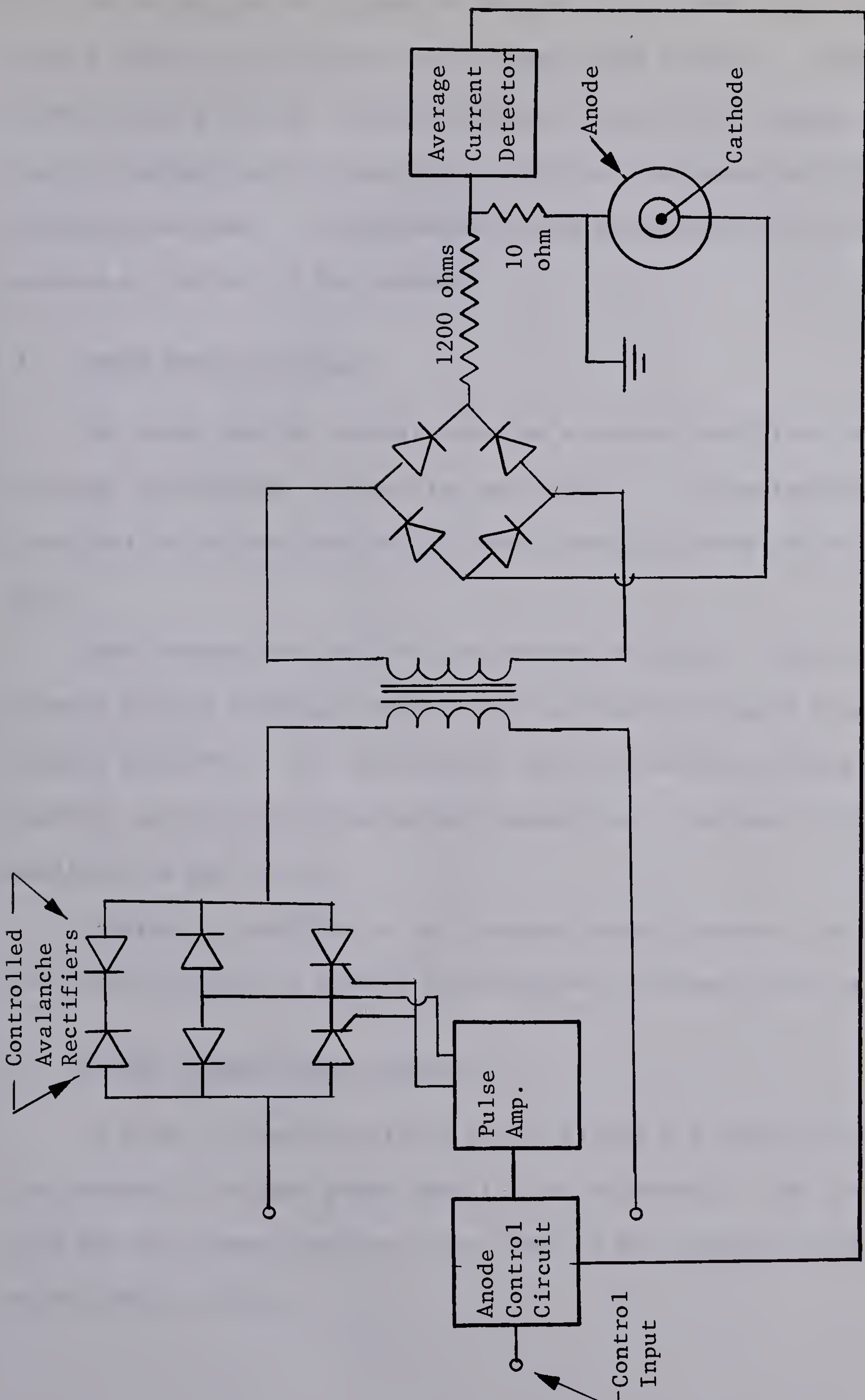
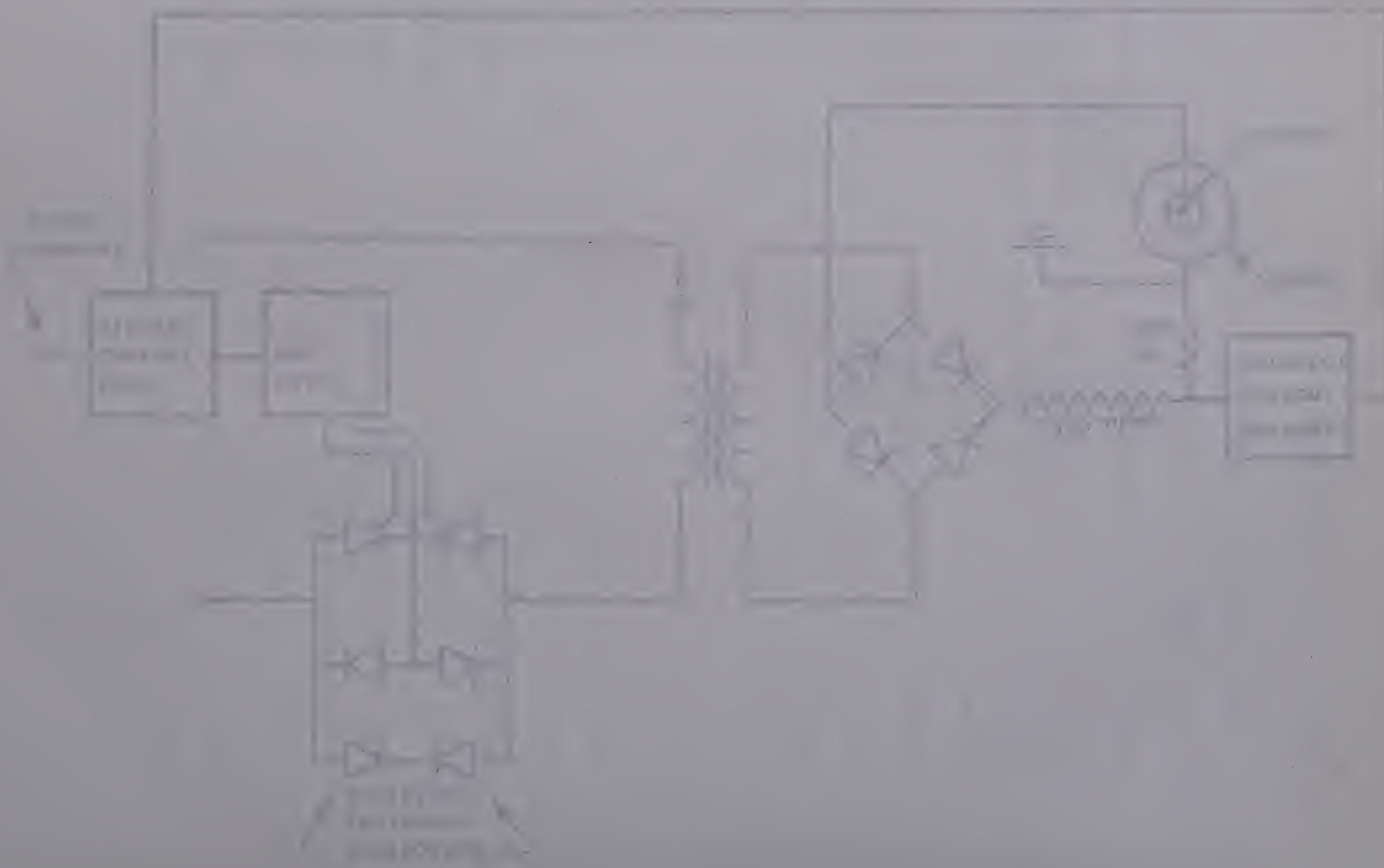


Fig. 3.1. Anode Supply Block Diagram.



The anode control system is designed as a closed loop system with feedback from a circuit which detects the average anode current. The system contains three separate blocks, the anode control circuit, the trigger pulse amplifier, and the average current detector. They are discussed individually in the following sections. A discussion of the performance of the complete system appears at the end of the chapter.

3.2 Anode Control Circuit

The anode control circuit contains a summing amplifier and a voltage controlled, synchronous, unijunction oscillator. The unijunction oscillator is identical to the one used in the heater control system and will not be described here.

Three inputs are used for the summing amplifier. One is used for the primary control input and another for the feedback signal from the average current detector. The third input, used for emergency shutdown of the anode current, is not used during normal operation. The gain of the summing amplifier is set at 5.6.

A circuit, identical to the average current detector, is used, preceeding the control input, to improve the transient response of the system.

3.3 S.C.R. Trigger Pulse Amplifier

In order to ensure reliable S.C.R. firing and stability of the anode control system, a trigger pulse amplifier is necessary. The circuit increases both the pulse power and the pulse width of the trigger pulses produced in the anode control circuit.

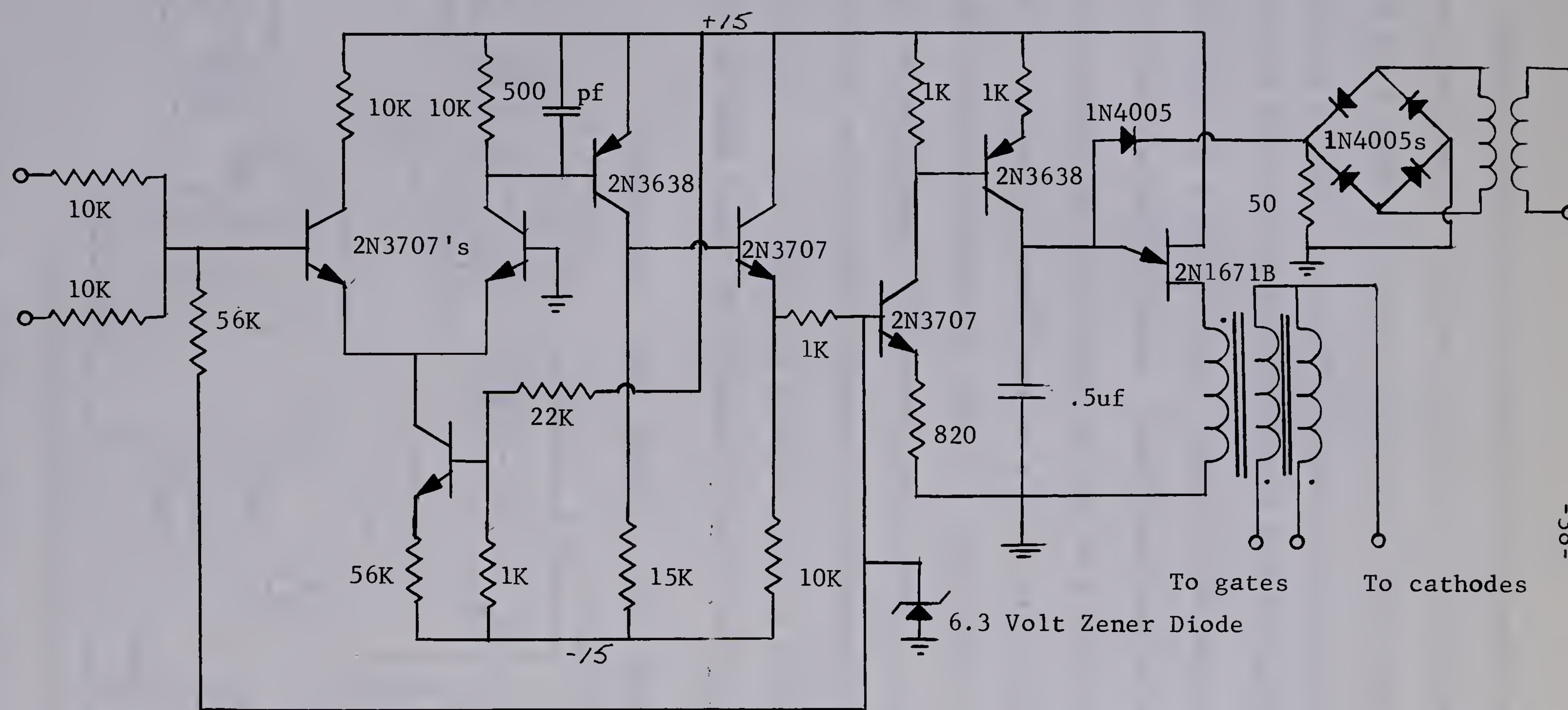


Fig. 3.2. Anode Control Circuit.

System instability could result if the narrow pulses of the anode control circuit are used to fire the S.C.R.'s. If, for example, the S.C.R.'s are fired for a conduction angle greater than that of the magnetron, S.C.R. conduction would probably not be established on the first trigger pulse. S.C.R. conduction would probably be initiated with the second trigger pulse, but this would result in a negative gain region in the anode transfer function.

A bistable multivibrator, transformer coupled to both the line voltage and the unijunction oscillator, is used as a pulse amplifier. The first pulse from the unijunction oscillator in a cycle will trigger the multivibrator. The multivibrator remains in this "on" state until the end of the line voltage cycle, at which time it is automatically reset. A saturating transistor is used as the output stage to provide the necessary gate power required to fire the S.C.R.'s. The output is directly coupled to the S.C.R.'s. A simple DC regulated power supply is included in the circuit.

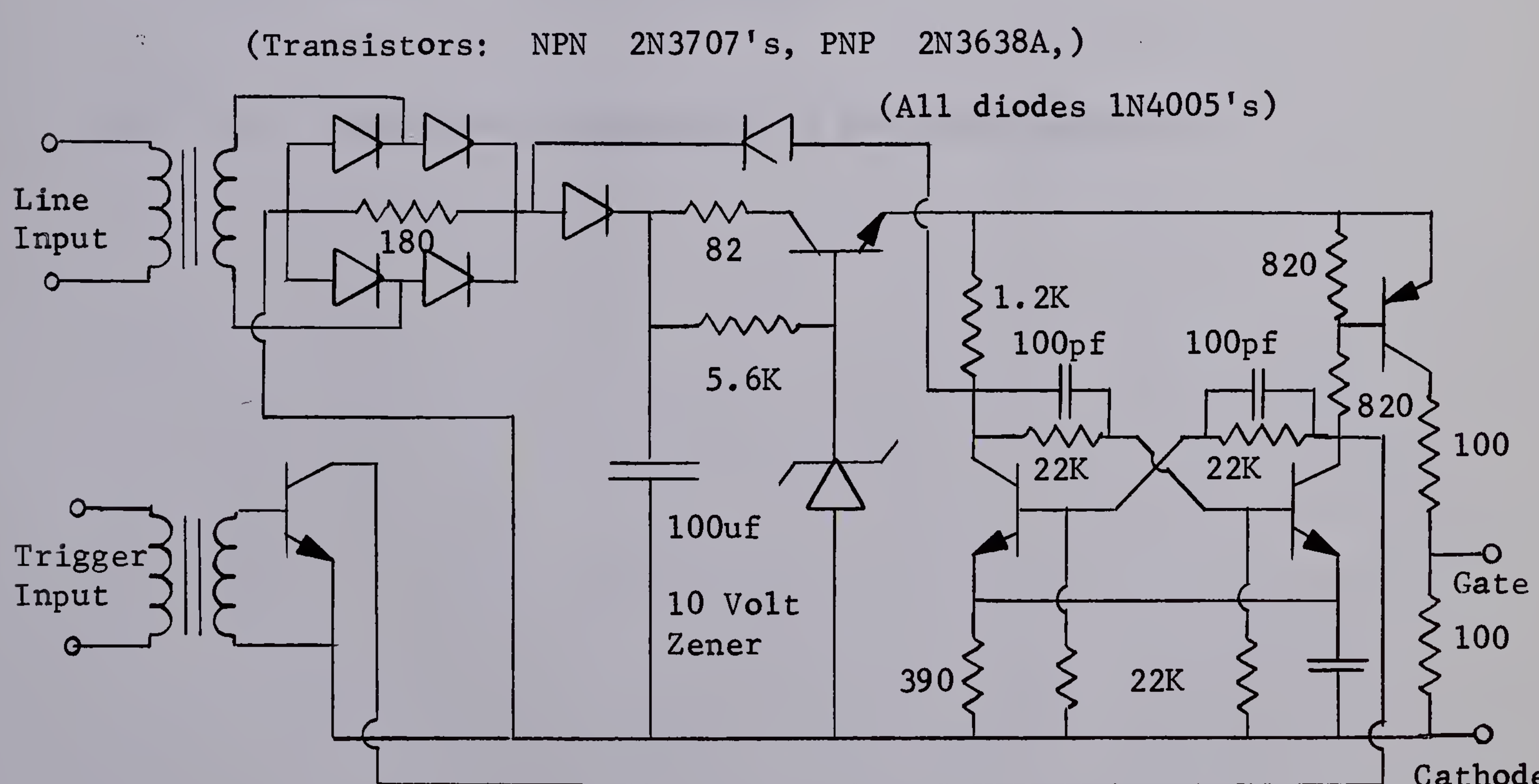
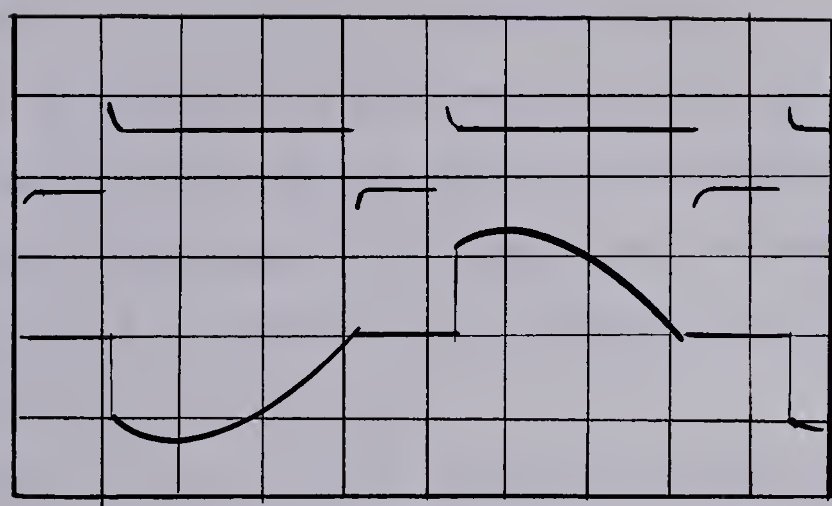


Fig. 3.3. S.C.R. Trigger Pulse Amplifier.



<u>Scale</u>

Horizontal -	2 Milliseconds per division.
Vertical	Uncalibrated.

Top Trace	Output of the pulse amplifier
Botton Trace	S.C.R. Waveshape.

Fig. 3.4. Waveshapes associated with the Pulse Amplifier.

3.4 The Average Current Detector Circuit

The function of the average current detector is to produce a DC voltage output which is proportional to the average anode current. The anode current is sampled by means of a 10 ohm, 20 watt resistor, placed in series with the magnetron, and arranged so that one end of it is at ground potential. A filter network is used to average the voltage across the resistor. A unity gain DC amplifier is used to provide a high impedance load for the filter and thereby isolate it from other circuitry.

The time constant of the filter is chosen as a compromise between low ripple and fast transient response. Experimental tests indicate that 200 milliseconds is a satisfactory value. Calculations, assuming a fullwave rectifed input, indicate that the ripple is less than 1% of the DC value.

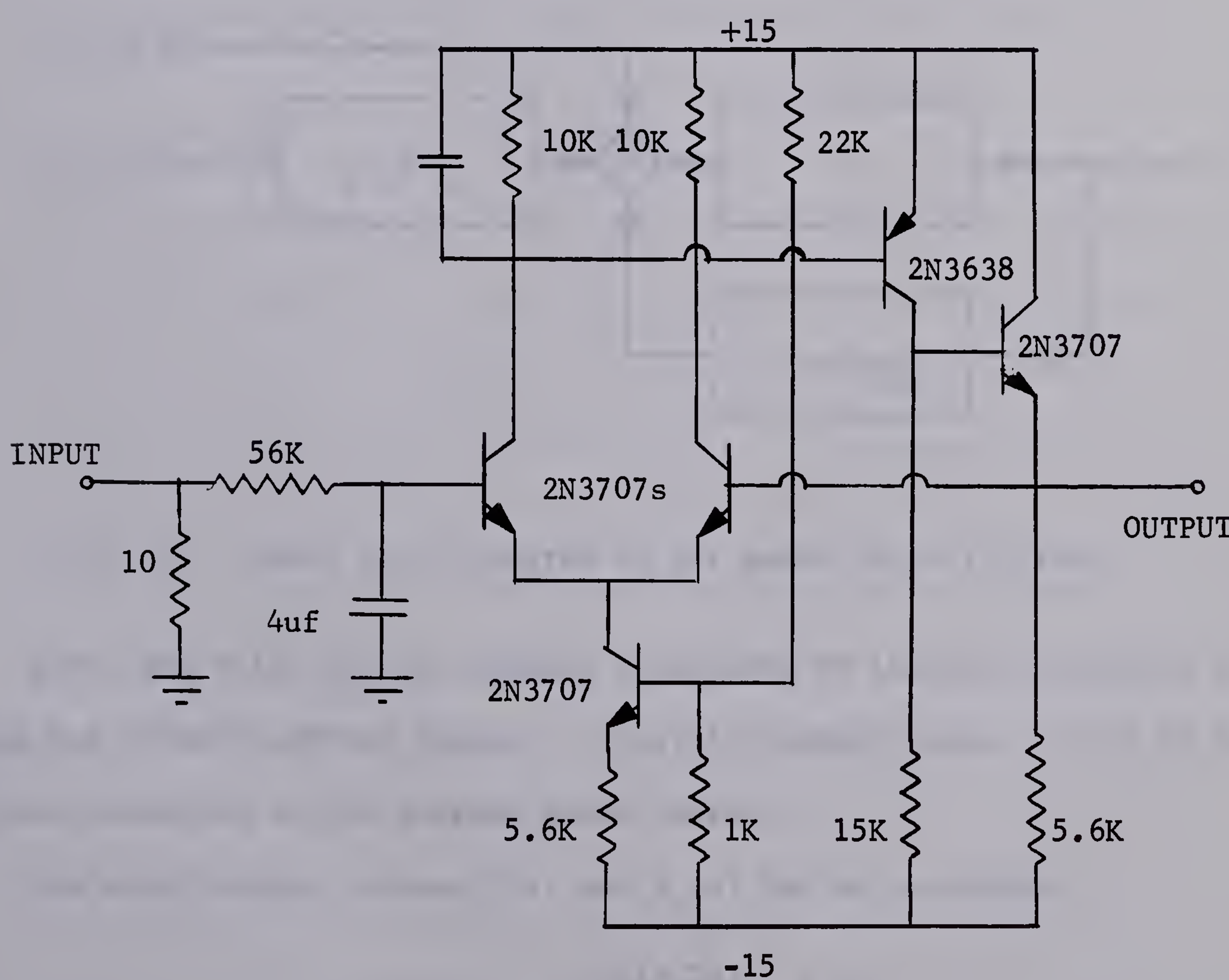


Fig. 3.5. Average Current Detector Circuit.

3.5 Performance of the Anode Control System

Important performance characteristics of the anode control system are its regulation to changes in line voltage, and its transient response. Exact prediction of these characteristics is difficult, especially for transient response, because the discontinuous nature of the S.C.R. system makes mathematical analysis cumbersome.

Regulation

The regulation of the anode system can be calculated assuming linear transfer functions for each of the circuits in the control system. Approximations yield the following servo block diagram.

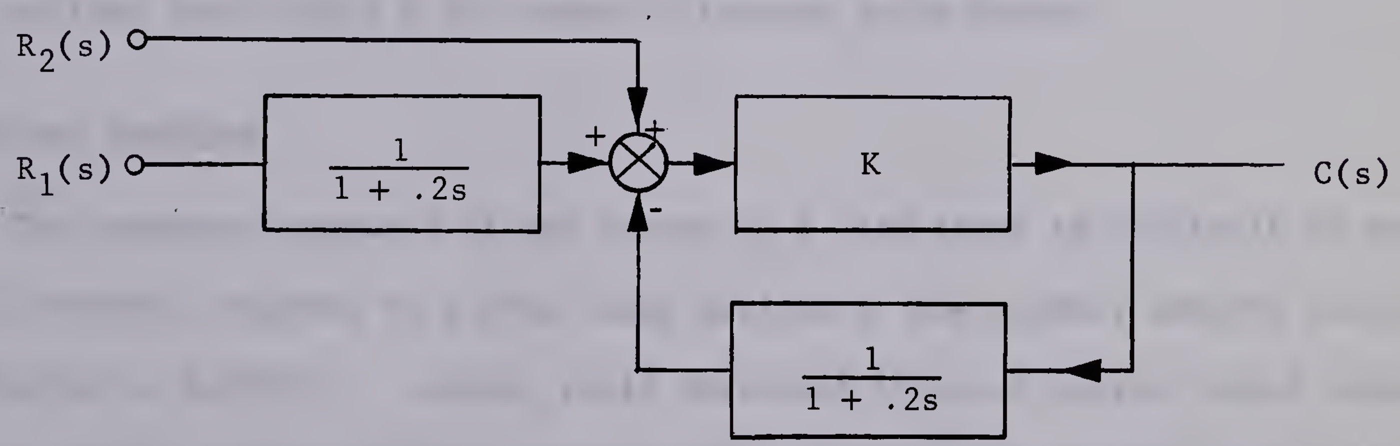


Fig. 3.6 Servo Block Diagram of the Anode Control System

$R_1(s)$ and $R_2(s)$ are the Laplace transforms of the input signals, $R_1(s)$ being the primary control signal. $R_2(s)$ is normally zero. $C(s)$ is the Laplace transform of the average anode current.

The relationship between $C(s)$ and $R_1(s)$ can be calculated.

$$C(s) = \frac{K(1+.2s)}{K + 1 + .2s} R_1(s) \qquad \ldots 3.1$$

If $R_1(s)$ is constant, the sensitivity of C(s) to changes in K, the open loop gain, can be calculated. The result is valid for steady state values only.

$$\frac{dC(s)}{C(s)} = \frac{1}{1+K}\frac{dK}{K}\,. \qquad \ldots\ 3.2$$

A change in line voltage can be related to a change in open loop gain by Eq. 1.6 of this thesis. For a magnetron half conduction angle of 52°

$$\frac{dK}{K} = 2.7\,\frac{dV}{V}\,. \qquad \ldots\ 3.3$$

The open loop gain of the system is 20. A 10% change in line voltage will therefore cause a 1.3% change in average anode current. This figure compares favourably with that of a saturable core reactor supply in which a 10% change in line voltage would yield a 5% change in average anode current.

Transient Response

The transient response of the system to a step input is difficult to predict. Fast transient response to a step input applied at the primary control input is not essential, however. Indeed, rapid increases in anode current would cause a sudden rise in cathode surface temperature. Because the thermal time delay between the heater element and the cathode surface is long, an immediate reduction of heater voltage would not prevent overheating of the cathode surface.

The compensation network used with the primary control input is designed, assuming a continuous system. Oscilloscope traces indicate that the system will respond to a step input in less than 1 second, which is certainly adequate for magnetron control.

Input $R_2(s)$ is used to provide immediate stoppage of anode current. A negative step input applied at $R_2(s)$ will stop the anode current in less than 4 milliseconds, the exact value depending on the time the step is applied in relation to the phase of the line voltage.

4. PROTECTION CIRCUITS

4.1 The Need for Protection Circuits

In order to protect the magnetron from adverse operating conditions caused by either a power supply failure or a human error, it is desirable to provide logic type control to override all other control circuits. The logic control serves to open the main contactor of the anode supply, and, if need be, to cut off the heater circuit.

The protection circuits are designed to shut off the anode supply if one of the following faults occurs.

(a) A failure of the heater control circuit caused by the light bulb burning out.

(b) Insufficient heater warm up time.

(c) Excessively high magnetron case temperatures which could be caused by a cooling system failure.

(d) Attempting to start the magnetron with anything but zero output power set on the anode control.

(e) An overload caused by high anode current.

(f) Accidental opening of the transformer enclosure.

In addition to shutting off the anode supply, the protection circuits will shut off the heater supply in the event that the light bulb used in the heater control circuit fails.

Lights are used to indicate the nature of the fault so that the necessary adjustments or repairs can be made. The logic control circuits are reset with push-button switches mounted on the main control panel.

4.2 Anode Protection

Figure 4.1 is a schematic diagram of the anode protection loop, showing the location of the various relays used.

The 180 second time delay relay prevents the application of high voltage to the anode before the heater has sufficient time to warm up. It is placed in the protection loop following the heater failure control to ensure that the heater is operating before the time delay period starts. It would otherwise be possible to start the anode supply immediately after correcting a heater failure, before allowing the necessary warm up period.

A manually operated switch is placed in the circuit after the time delay for use as an on-off switch.

A microswitch, driven by a cam mounted on the anode control potentiometer, is used to ensure that the pot is set to zero when the contactor is closed. Otherwise it would be possible to have the magnetron start at full power when the contactor is closed. A relay acts to override the microswitch when the contactor is closed.

Switching of the relays is accomplished with logic type circuits. In each case a Schmitt trigger circuit is combined with a bistable multivibrator. The relays are used as load resistors in the multivibrator circuits.

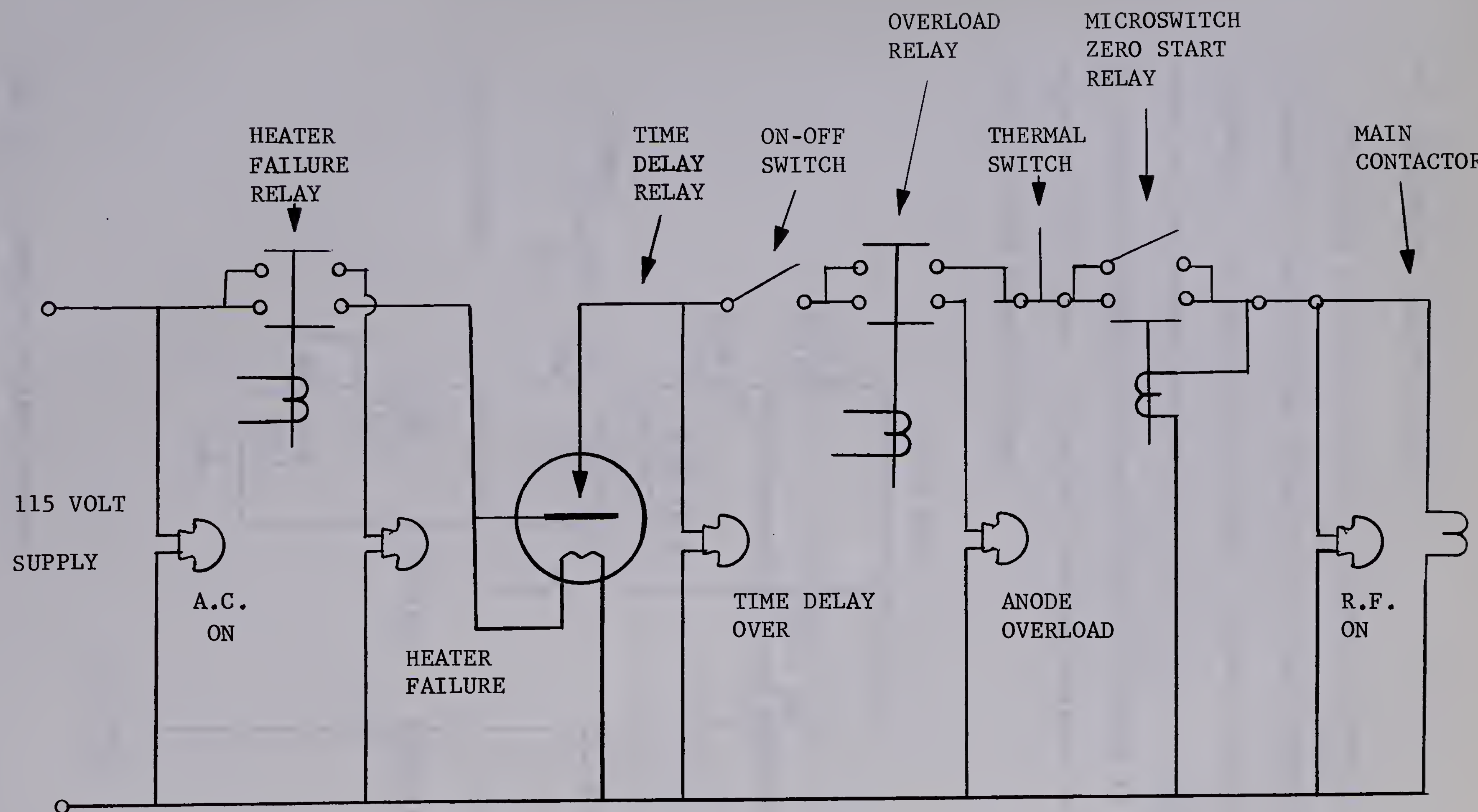
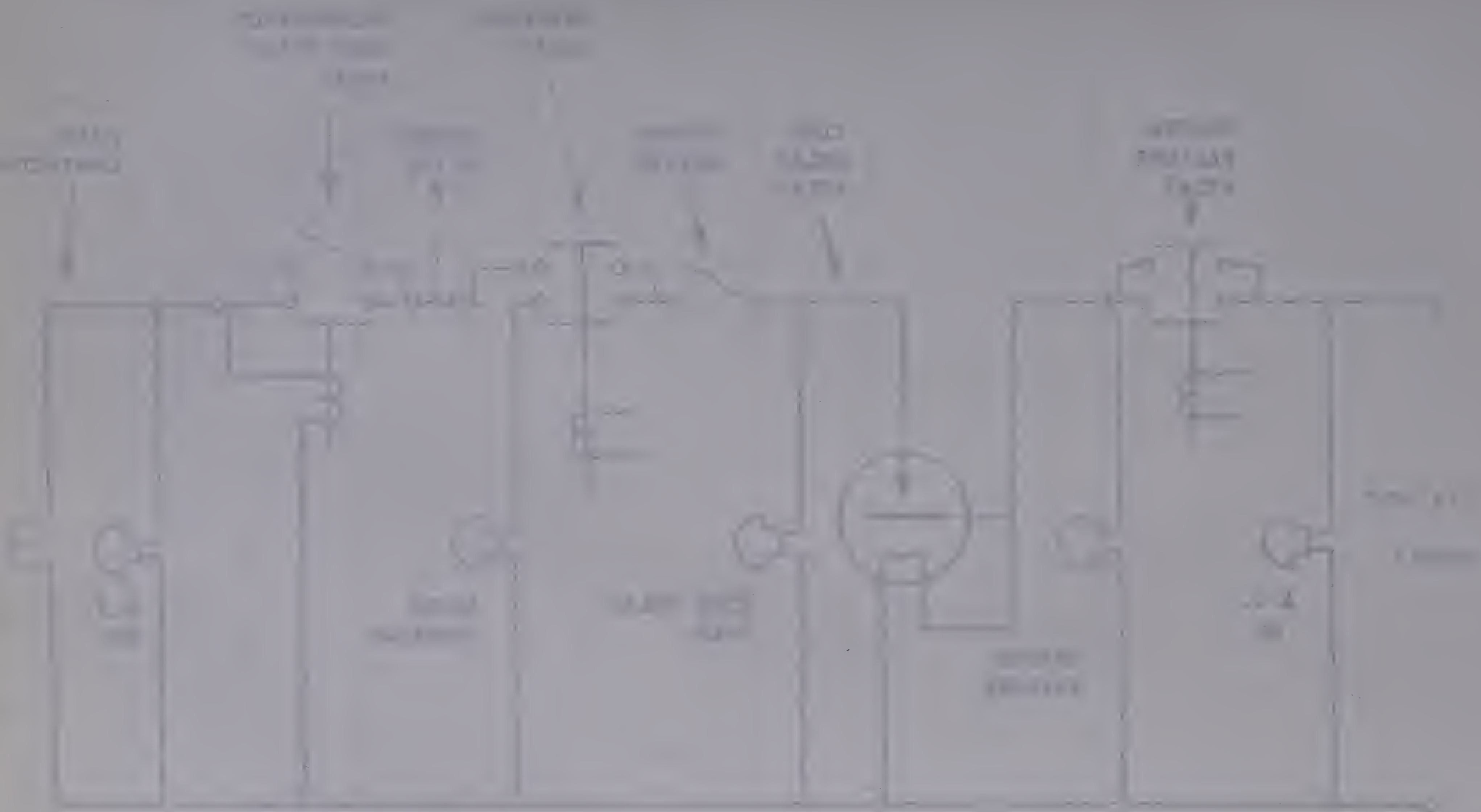


Fig. 4.1. Anode Protection Loop.



4.3 The Anode Overload Circuit

The anode overload circuit is designed to shut off the anode current if it exceeds a predetermined value in the range between 300 milliamps to 1000 milliamps. The input to the Schmitt trigger circuit is obtained from the output of the average current detector circuit. The Schmitt trigger circuit is therefore designed for fire with input voltages in the range between 3.0 volts and 10.0 volts. A variable emitter resistor is used to set the firing voltage of the Schmitt trigger.

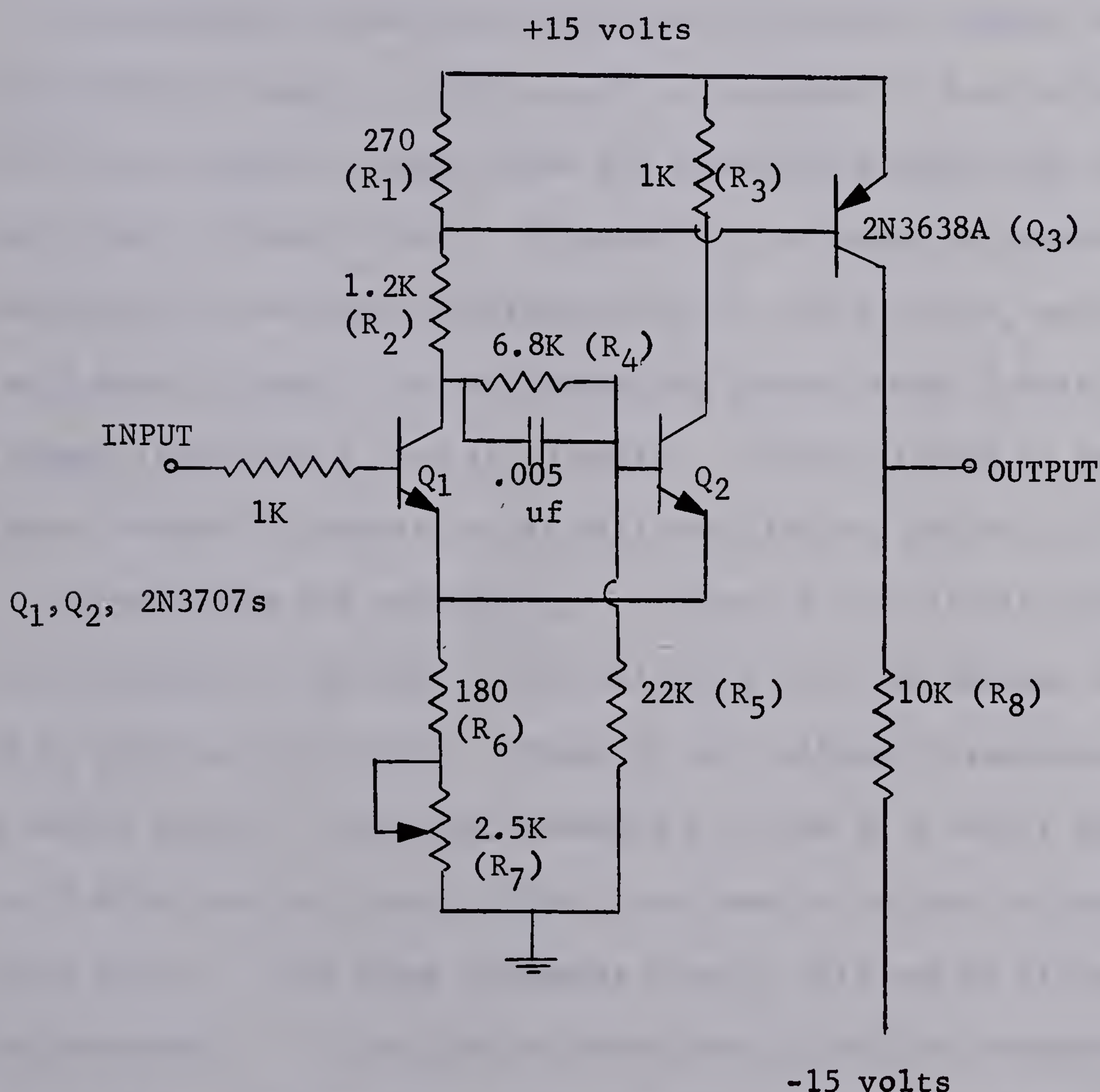


Fig. 4.2. Schmitt Trigger Circuit.

In order for the Schmitt trigger to operate it is necessary to make $R_1 + R_2 > R_3$. Generally, $R_1 + R_2$ is made only slightly larger than R_3 to minimize hysteresis effects; however, in this application it is only necessary to have the Schmitt trigger circuit fire; it can reset itself at any voltage above zero volts.

The biasing resistors R_4 and R_5 are chosen so that Q_2 will be saturated with V_F (input voltage to fire the circuit) set at its two extremes. The circuit is designed so that it will work with transistors which have an $h_{FE} > 40$.

An amplifier stage is used between the Schmitt trigger and the bistable multivibrator stages. This circuit is designed to have an output of -15 volts before the Schmitt trigger fires and something greater than +5 volts after the Schmitt trigger fires. In order for the stage to operate properly, it is necessary for the output voltage to be -15 volts with V_F set at 3 volts, before the Schmitt trigger fires and something greater than -5 volts after the Schmitt trigger fires with V_F set at 10 volts. If the circuit is designed to meet these extremes in operation, it will work for any setting of V_F.

Considering the extremes in operation of the circuit, with V_F set at 3.0 volts and the circuit in its unfired state, the emitter to base voltage of Q_3 will be 0.377 volts. Since Q_3 is a silicon transistor, it will be in a cutoff state. Similarly, assuming Q_3 to be in a cutoff state with V_F set to 10 volts and the circuit fired, the base to emitter voltage of Q_3 will be 0.933 volts. This value indicates that Q_3 will not be off as assumed, but will be conducting. It can also be shown that Q_3 will be saturated providing $h_{FE_3} > 1.5$. The amplifier will therefore work properly for any setting of V_F between 3.0 volts and 10.0 volts. The output in the unfired state will be -15 volts. The output in the fired state will be +15 volts minus the collector to emitter saturation voltage of Q_3.

4.4 The Bistable Multivibrator Circuit

The bistable multivibrator is designed to power a 2500 ohm DC relay which requires a holding current of 8 milliamps. Since 20 volts is needed to operate the relay, the circuit is designed to operate between +15 and -15 volts. An emitter resistor is included in the circuit so that the emitters will operate at -11 volts. The relay is arranged in the circuit so that it is held on during regular operation, and is shut off during an overload.

The circuit is designed so that the transistors will be in either the saturated or cutoff state. Resistors are chosen so that the circuit will operate with transistors having an $h_{FE} > 30$.

Triggering of the circuit is accomplished by means of a positive pulse applied to the base of Q_4. A diode is used to prevent the circuit from resetting when the Schmitt trigger resets. A push-button switch, connected between the base of Q_4 and -15 volts, is used to manually reset the multivibrator.

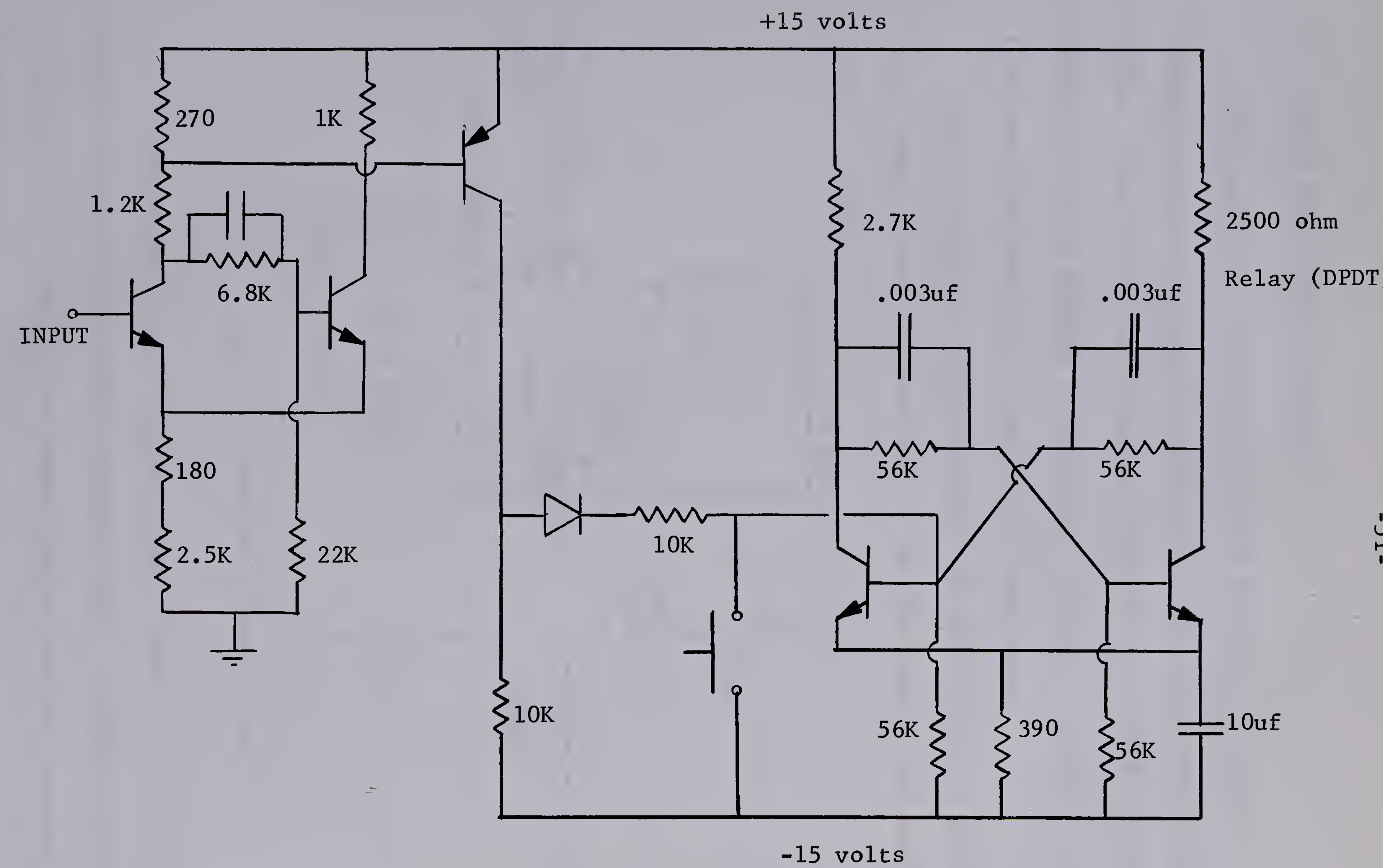


Fig. 4.3. Complete Anode Overload Control Circuit.

4.5 <u>Light Bulb Failure Circuit</u>

The light bulb failure circuit is designed to shut off both the heater voltage and anode current if the light bulb used in the heater control system fails. The circuit is triggered whenever the output voltage of the r.m.s. detector circuit exceeds 9.0 volts. A filter is used to prevent the circuit from being accidentally triggered by transients.

Only the Schmitt trigger portion of the circuit is shown below, the multivibrator stage is identical to the one used in the anode overload control circuit.

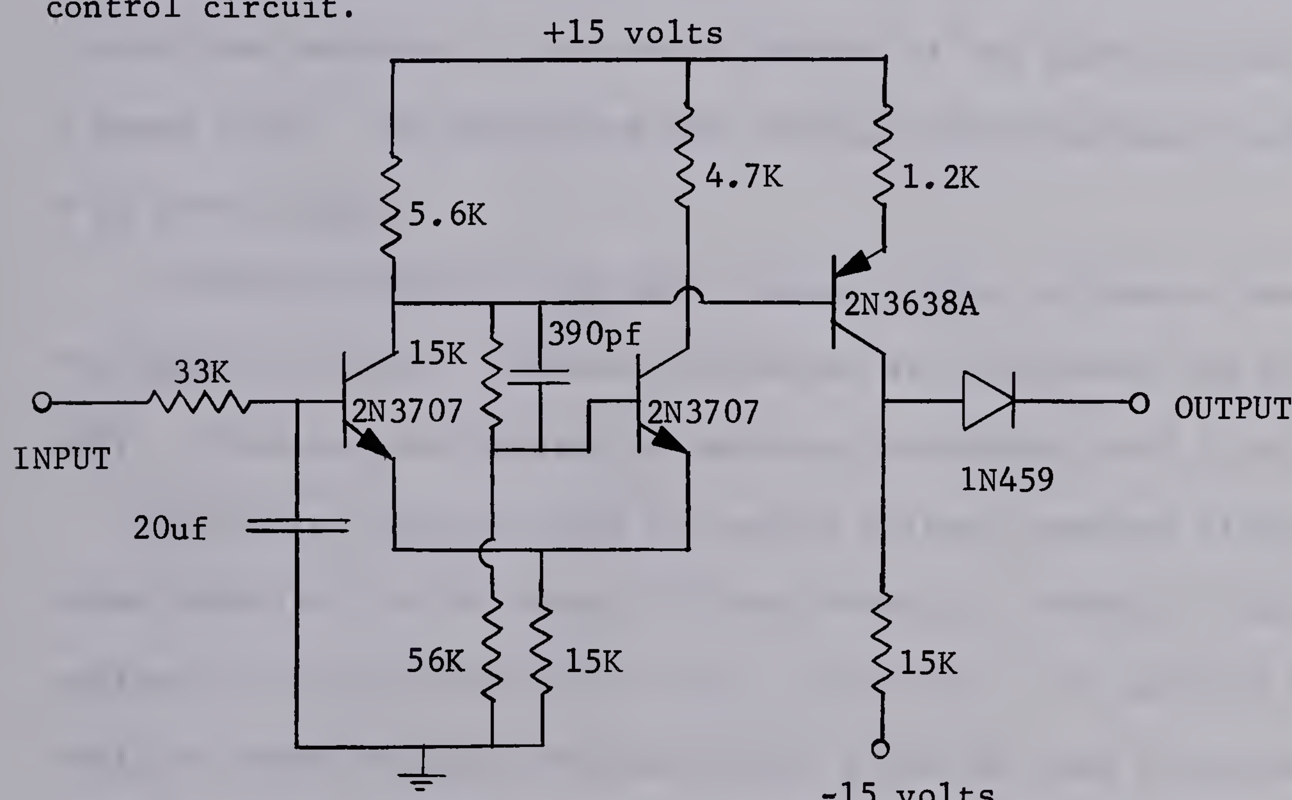


Fig. 4.4 Schmitt Trigger Circuit

A double pole, double throw relay is used in the multivibrator stage. Besides breaking the anode protection loop, it shorts out the trigger pulses to the S.C.R.'s used in the heater control system, thereby stopping the heater voltage.

5. TESTS AND RESULTS

A series of tests were carried out to evaluate the performance of the heater control system and the anode control system. The tests were carried out in several stages. In initial tests dummy loads were substituted for the magnetron and inputs to the control systems were simulated.

5.1 Heater Control System

The relationship between heater voltage and the average value of anode current was measured. The heater element of the magnetron was replaced with a dummy load. The input from the average current detector was simulated with a DC power supply.

A Hewlett Packard, type 3400, thermal r.m.s. voltmeter was used to measure the heater voltage. Ordinary voltmeters are calibrated for sine wave inputs only. They are inaccurate for measuring waveshapes with high crest factors.

The tests indicated that the heater voltage remained within 1% of the value specified in the heater voltage schedule. However, this result is subject to the accuracy of the r.m.s. voltmeter. It can only be concluded that the heater voltage remained within a few per cent of the desired value but certainly within the +5% and -10% range specified for the 7292 magnetron.

An attempt was made to measure the regulation of the system to changes in line voltage. For a 10% change in line voltage, no change in heater voltage was noted.

5.2 Anode Control System

The anode control system was first tested with a dummy resistive load. The tests proved successful. Operation was then attempted with an Amperex DX-260, one kilowatt magnetron. The S.C.R.'s were placed in the primary circuit of the anode transformer. Resistive current limiting was used.

The magnetron was operated successfully with anode currents in the range between 100 milliamps and 375 milliamps or full power.

An attempt was made to operate the magnetron with anode currents below the specified minimum of 100 milliamps. It was hoped that the magnetron could be operated at a very low output power and thereby be used for experiments requiring a low power generator. Anode currents as low as 2 milliamps were obtained with good stability of the control system.

Mechanical vibrations were observed at these low values of anode current, however. No conclusions were reached as to the origin of the vibrations. Because the magnetron was operated below the minimum value of anode current recommended for the tube, it is thought that the vibrations could have originated in the tube itself. It was concluded that the tube was unsuitable for operation at very low levels of output power.

5.3 Tests with the 7292 Magnetron

Initially, only the heater control system was tested. The anode current was controlled with a variac mounted at the primary of the anode transformer.

High frequency oscillations superimposed on the anode current were noted, particularly at the start and end of magnetron conduction. Further examination showed that the oscillations also occurred in the voltage across the bridge rectifier. No conclusions were reached as to whether the oscillations originated

in the bridge rectifier or in the magnetron itself.

The oscillations were unexpected as they had not appeared in tests of a saturable reactor controlled supply. The oscillations were attributed to the use of resistive current limiting. It is thought that the use of inductive current limiting would limit the oscillations.

Careful examination of the oscillations with an oscilloscope indicated that the frequency of the oscillations was primarily 10 MHz although higher frequency components were noted. Because of the high frequency and the length of the leads connecting the magnetron, transformers, S.C.R.'s, and the control system, the oscillations appeared in all control circuitry. Line interference was also noted and the use of interference filter is recommended.

A. Heater Control System

The linearity of the relationship between the heater voltage and the anode current was measured. The effects of the oscillations were evident at anode currents below 100 milliamps. The non-linearity is illustrated in Fig. 5.1. Because the heater voltage was still within the desired limits and because the oscillations were not random, occurring only at the start and end of magnetron conduction, it was decided to accept the performance of the heater control system. Actually, the non-linearity does not occur within the normal operating range of the magnetron. The minimum recommended anode current is 100 milliamps.

The transient response of the heater control system was measured. A step change was made in the anode current, by switching the magnetron on or off, and the heater voltage was recorded. The Hewlett Packard voltmeter, used to measure the heater voltage, provides a DC output which was used to drive a strip chart recorder. This method is not very accurate, however, because

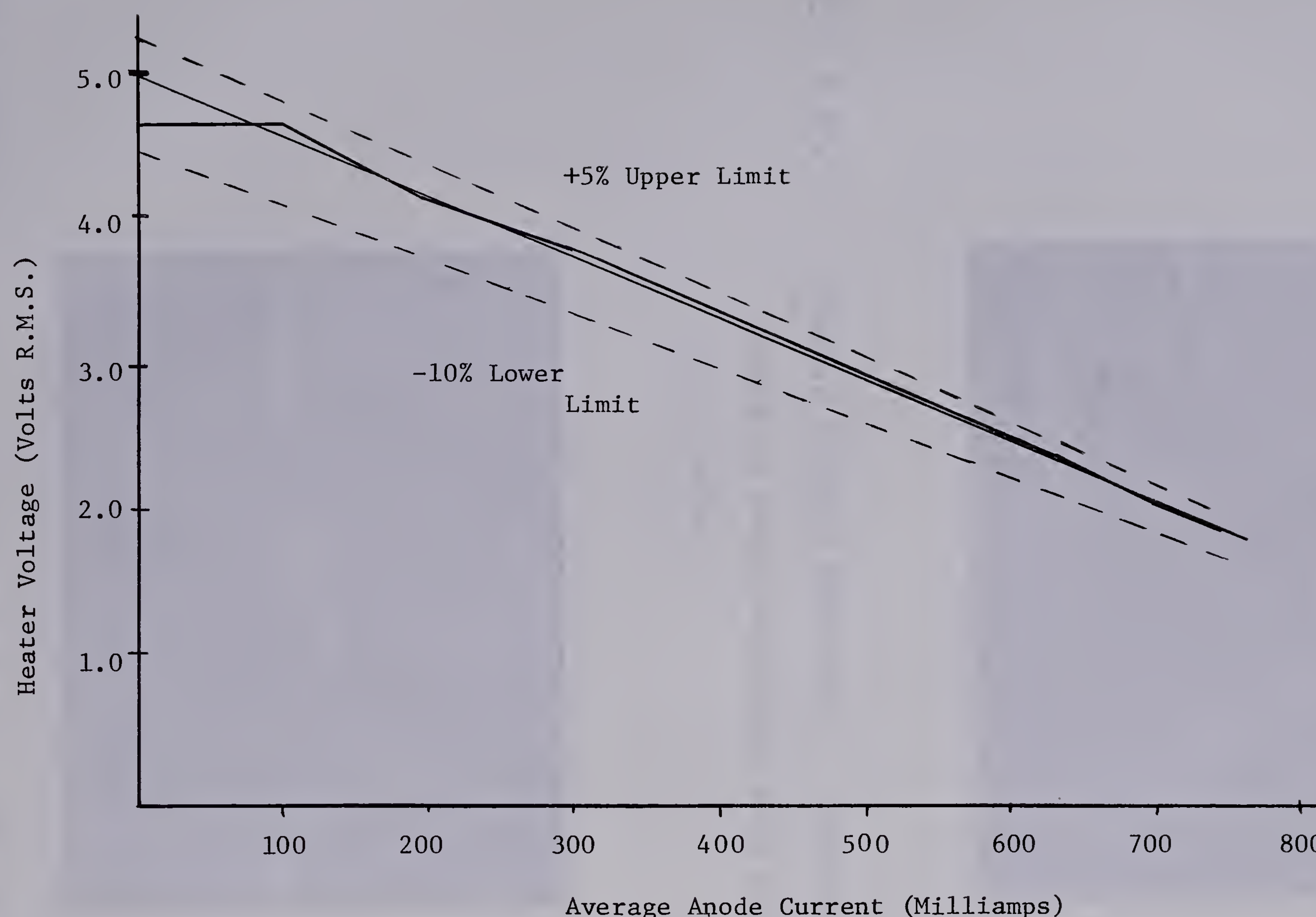


Fig. 5.1 Measured Relationship Between Heater Voltage (R.M.S.) and Anode Current (Average)

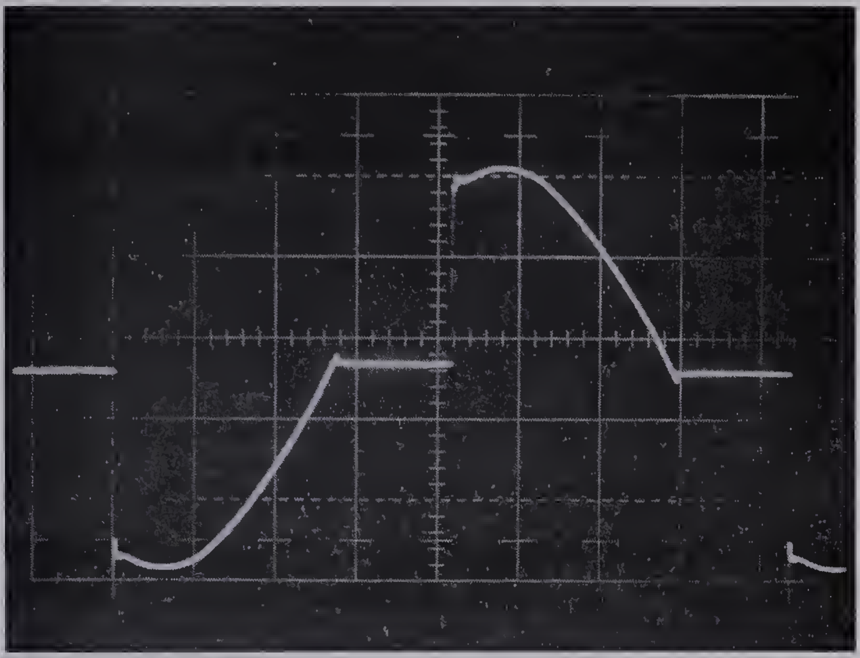


Fig. 5.2(a) Heater Voltage Waveshape. (Observed at the input to Power Amplifier with anode current set at zero.)

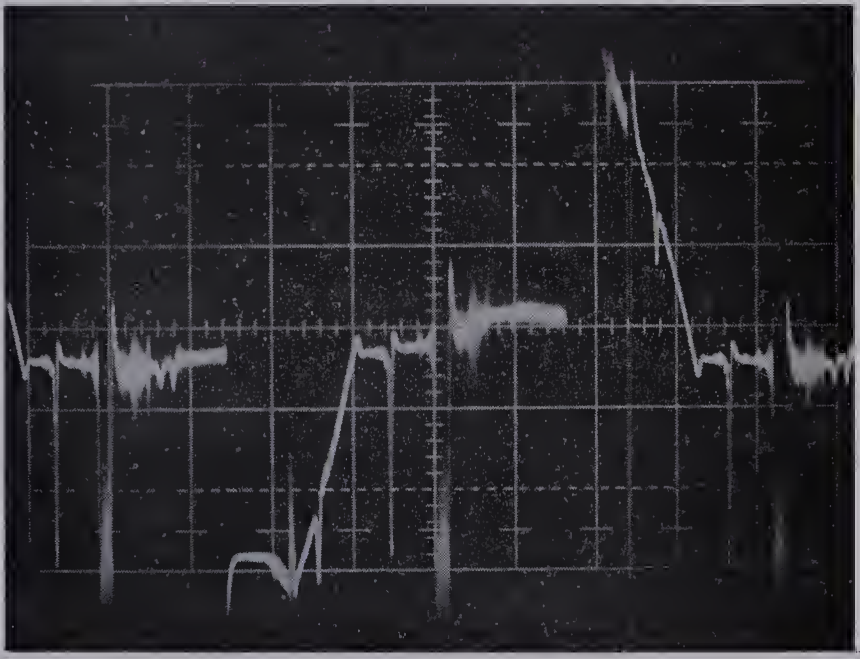
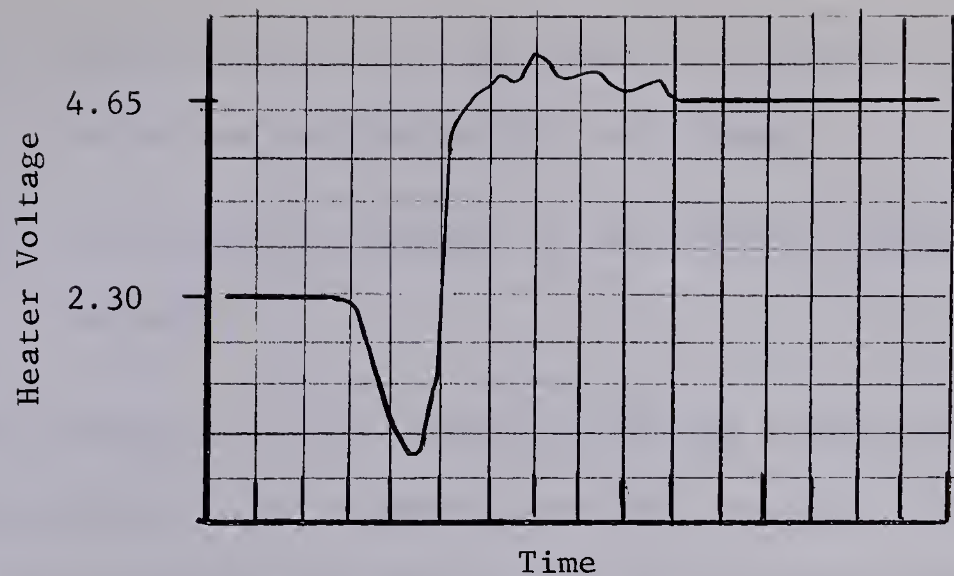


Fig. 5.2(b) Heater Voltage Waveshape. (Observed at the input to Power Amplifier with anode current set at 650 milliamps.)



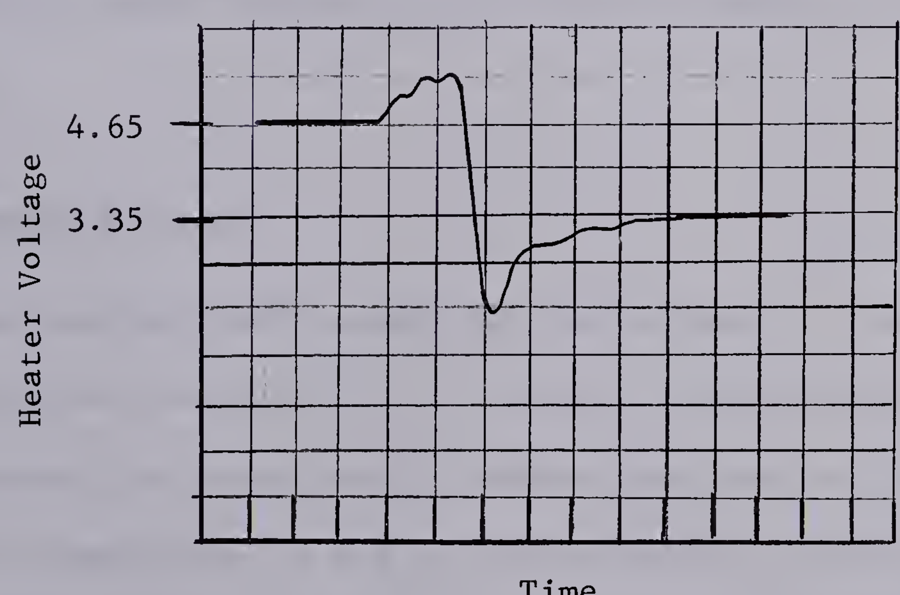
Scale

Horizontal 1 second per division

Vertical Heater Voltage (uncalibrated)

Fig. 5.3(a) Heater Voltage Transient Response.

(Anode Current Switched From 650 Milliamps To 0 Milliamps)



Scale

Horizontal 1 second per division

Vertical Heater Voltage (uncalibrated)

Fig. 5.3(b). Heater Voltage Transient Response.

(Anode Current Switched From 0 Milliamps To 400 Milliamps.)

(a) The DC output of the voltmeter is proportional to the meter deflection and the scale is not linear.

(b) The response is subject to the transient response of the voltmeter.

The results (Fig. 5.3) indicate that the heater voltage will settle to its steady state value within about five seconds. Because the saturable core reactor controlled supply requires up to three minutes to settle within the +5% and -10% limits, the transient response was considered satisfactory. The results are perhaps unrealistic, however, because the important time constant in the system is the time required for the cathode surface temperature to obtain its steady state value. It is believed that the thermal time delay between the heater element and the cathode surface is much longer than five seconds.

B. Complete System

The complete power supply was tested next. The oscillations present in the system had no effect on the anode control system. This is understandable because the anode control system functions to fire the S.C.R.'s and once this is accomplished it has no further control over the magnetron until the next cycle. The oscillations can only occur after the S.C.R.'s have fired.

The protection circuitry was tested and found to be satisfactory. The anode overload control was tested in the range between 300 milliamps and 750 milliamps anode current.

The magnetron was operated over the full range of allowable currents. Even below the minimum value of anode current no vibrations were noted.

6. DISCUSSION

6.1 General

The purpose of this thesis was to investigate the use of solid state control systems in a power supply for low power continuous wave magnetrons. A supply was designed and built specifically for the Philips 7292 magnetron.

It was found that the transient response and regulation of the control systems were much improved over those of comparable saturable core reactor controlled supplies. Weight and size were also reduced.

The most significant advantage of the solid state control system was found to be the ease with which the output power of the tube could be controlled. A low voltage signal is used to control the anode current. Such control would be of particular advantage in multiple tube systems, or in systems subject to intermittent loading.

In mutliple tube systems there is a danger of energy from one tube being coupled into another tube. Ferrite isolators are not considered an answer to the problem because of their extremely high cost.

The reverse flow of power into a microwave tube can be detected electronically. A logic control system combined with the solid state control system could be used to stop one or more tubes if cross coupling occurred. The solid state system would be capable of stopping any tube in the system within four milliseconds.

In microwave systems where loading is intermittent it is desirable to stop the generator as soon as the load is removed. The reasons are twofold. Reflected power from an unloaded cavity can quickly destroy a tube. Operating the cavity without a load is also uneconomical. The solid state

control system can easily be programmed to start and stop the generator as the loading conditions change.

The systems presented in this thesis, although designed specifically for one magnetron, could easily be adapted to other low power tubes. In some cases, however, other types of solid state control systems are to be desired. They are described briefly in the following sections. Because the heater control system and the anode control system are essentially separate systems they will be described separately.

6.2 Heater Control System.

The heater control system is particularily important to magnetrons at the present time. The reasons are mainly economic. The life of present tubes is considered insufficient for many industrial applications.[13] Typically it is the cathode and the cathode heater which fail. Maximum life of the thermionic type cathodes can only be achieved if the cathode surface is operated at or near its proper temperature.

The cathodes of magnetron type tubes are subject to extensive electron back bombardment. The heating effect of this back bombardment is sufficient to warrant control of the heater voltage. Electron back bombardment is a function of two operating variables, the output power of the tube, and the loading conditions of the tube. Electron back bombardment increases as the standing wave ratio of the load increases.

In order to provide accurate control of the cathode temperature it is necessary to continuously monitor the temperature. This is extremely difficult with most tubes, mainly because the cathode is enclosed in an

evacuated space. An indirect method was attempted with the Philips 7292 magnetron.

The 7292 magnetron has a glass window mounted in the anode block which permits visual inspection of the cathode. The quantity of light emitted by the cathode is highly dependent of the surface temperature.

An LS-400 photo transistor was mounted so as to measure the quantity of light emitted by the cathode. The device was included in a circuit which provided an output voltage which was proportional to the light emitted.

To test the sensitivity of the detector the magnetron was operated with its conventional power supply at different values of anode current and the output of the detector was measured.

Even though the heater voltage was maintained at its proper level for all values of anode current at which tests were made, large variations in the output of the detector were noted. The results were repeatable.

A system could therefore be designed so as to adjust the heater voltage in order to regulate the output of the detector. Such a system would then provide close regulation of the cathode temperature.

There are problems associated with such a system however. The system would be highly dependent on the optical properties of the window. There are reasons to believe that these change during the operating life of the magnetron. The window of a used magnetron appears cloudy when compared to that of a new magnetron.

Alignment of such a system is another problem. The system would be useless if it were regulating the cathode temperature at a value other than the proper value.

Although cathode temperature control is a real problem at the present time, future developments may eliminate the problem completely. Raytheon[14] has developed a cold cathode which has been sucessfully operated in a high power tube. The cold cathode, unlike present thermionic cathodes, requires no heater power. In the cold cathode emission is stimulated by electron back bombarment.

6.3 Anode Control System.

Magnetrons can be divided into two classes, those with fixed field magnets, and those with variable field magnets. The Philips 7292 magnetron and the Amperex DX-260 magnetron belong to the first class.

In the fixed field class of magnetrons, the output power of the tube is controlled by varying the anode current. Two methods are commonly used. Some supplies use a fixed input voltage but vary the current limiting device impedance to obtain control. Other supplies use a fixed value of current limiting impedance but vary the input voltage. The supply discussed in this thesis used the latter method. With either method the control system must be able to handle the full power delivered to the magnetron. This is often awkward with high power tubes.

Recently introduced tubes are of the variable field type. The Litton L-3821, 2.5 kilowatt tube is an example. If the input voltage and the series impedance are fixed, the output power is dependent on the current delivered to the electromagnet. This current is in the order of one ampere at full power.

It is felt that control of such a magnetron would be much easier than of magnetrons of the fixed field type. Solid state control systems could again be used.

CONCLUSIONS

When compared to present saturable core reactor controlled supplies the solid state controlled supply offers the following advantages.

(1) Much faster transient response.

(2) More accurate control of the heater voltage.

(3) Better regulation to line voltage changes.

(4) Voltage control of the output power of the magnetron permitting programmed operation and the use of logic type control.

(5) Reduced size and weight.

Advances in magnetron design in the future may make the control systems discussed in this thesis obsolete. The introduction of the cold cathode may eliminate the heater control system completely. In variable field type magnetrons the output power will be controlled by varying the magnetic field rather than the anode current directly.

APPENDIX I

This appendix contains photocopies of specification sheets and operating notes for three different low power continuous wave magnetrons.[15, 16, 17] They are the Philips 7292 magnetron, the Amperex DX-260 magnetron, and the Philips L-3821 magnetron.

TYPICAL OPERATION AT 2.0 KW POWER OUTPUT[1]

ABSOLUTE RATINGS

Average Anode Current	0.8 Amp max.
	0.1 Amp min.
VSWR	
Between 0.37λ and 0.44λ (reference plane A)	4.0 max.
Remaining Region	5.0 max.

TYPICAL OPERATION

Average Anode Current[7]	0.75 Amp
Peak Anode Current	2.0 amps
Anode Voltage[8,9]	4.6 kV
VSWR	
Between 0.37λ and 0.44λ toward Load from Reference Plane A (phase of sink)	3.0 max.
Remaining Region	3.0 max.
Output Power[9]	2.0 kW
Efficiency[9]	55 %

See Figures 4 and 5 for further information

OPERATING NOTES

Anode Supply

The magnetron should be operated from an unfiltered anode supply with a single phase full wave, two phase half wave or three phase half wave rectified unfiltered system. Operation of the magnetron from a filtered supply generally results in a lower magnetron efficiency.

Heater Voltages

The heater voltage must be reduced immediately after applying the anode power to compensate for additional heating of the cathode when the magnetron is oscillating. In Figure 7, the reduced heater voltage is given as a function of average anode current.

The longest operating life of the magnetron will be obtained if the heater voltage is reduced to a value as given by the fully drawn line A. The heater voltage should be adjusted within +5% and -10% as given by the dashed lines which border the hatched area.

When the microwave heating equipment is designed for a predetermined number of steps of output power level, the reduced heater voltage for each step must be set to a value within the area bordered by the curves B and C and preferably in such as way that it is within or close to the hatched area. Under no circumstances (for instance with line voltage variations) should the heater voltage exceed the limits given by the curves B and C.

ELECTRICAL

Cathode[6]	dispenser type, indirectly heated ac or dc
Anode Current[7]	0.750 Amp
Anode Voltage[8, 9] at an Average Anode Current of 0.750 Amp	4600 ± 200 Volts
Heater Starting Voltage[10]	5.0 + 5% - 10% Volts
Heater Current at 5.0 Volts	35 Amps
Cold Heater Resistance	0.02 ohm
Output Coupling System	53.5 ohms 1-5/8 coaxial

ABSOLUTE MAXIMUM RATINGS

Heater Starting Voltage	5.25 Volts
Heater Surge Current (peak value of wave shape)	140 amps
Average Anode Current[7]	0.9 Amp
Peak Anode Current[11]	2.1 amps
Anode Temperature [3]	125°C
Cathode Radiator Temperature[3]	180°C

TYPICAL OPERATION AT 2.5 KW POWER OUTPUT[1]

To operate at this power level, an impedance transformer section having the following characteristics with respect to reference plane A on outline drawing must be placed between magnetron and load: (See accessories.)

VSWR = 1.5
Phase Position ± 0.41 (phase of sink region, see Figures 5 and 6)

ABSOLUTE RATINGS

Average Anode Current	0.9 Amp max. 0.1 Amp min.
VSWR	
Between 0.37λ and 0.44λ (reference plane A')	2.5 max.
Remaining Region	4.0 max.

TYPICAL OPERATION

Average Anode Current[7]	0.85 Amp
Peak Anode Current	2.0 amps
Anode Voltage[8, 9]	4.6 kV
VSWR	
Between 0.37λ and 0.44λ toward Load from Reference Plane A' (phase of sink)	2.5 max.
Remaining Region	2.5 max.
Output Power[9]	2.5 kW
Efficiency[9]	60 %

See Figures 4 and 6 for further information

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[6] Before the application of high voltage the cathode must be heated for at least 2 minutes.
[7] Measured with a moving coil instrument (average value).
[8] Measured with filtered dc at an air gap field of 1100 Gauss.
[9] At a matched load (VSWR less than 1.05).
[10] When the magnetron is oscillating, the heater voltage must be lowered immediately. For specific values see "Operating Notes".
[11] The dynamic internal impedance of the anode voltage supply circuit must be adjusted so that the peak anode current is not exceeded during operation.

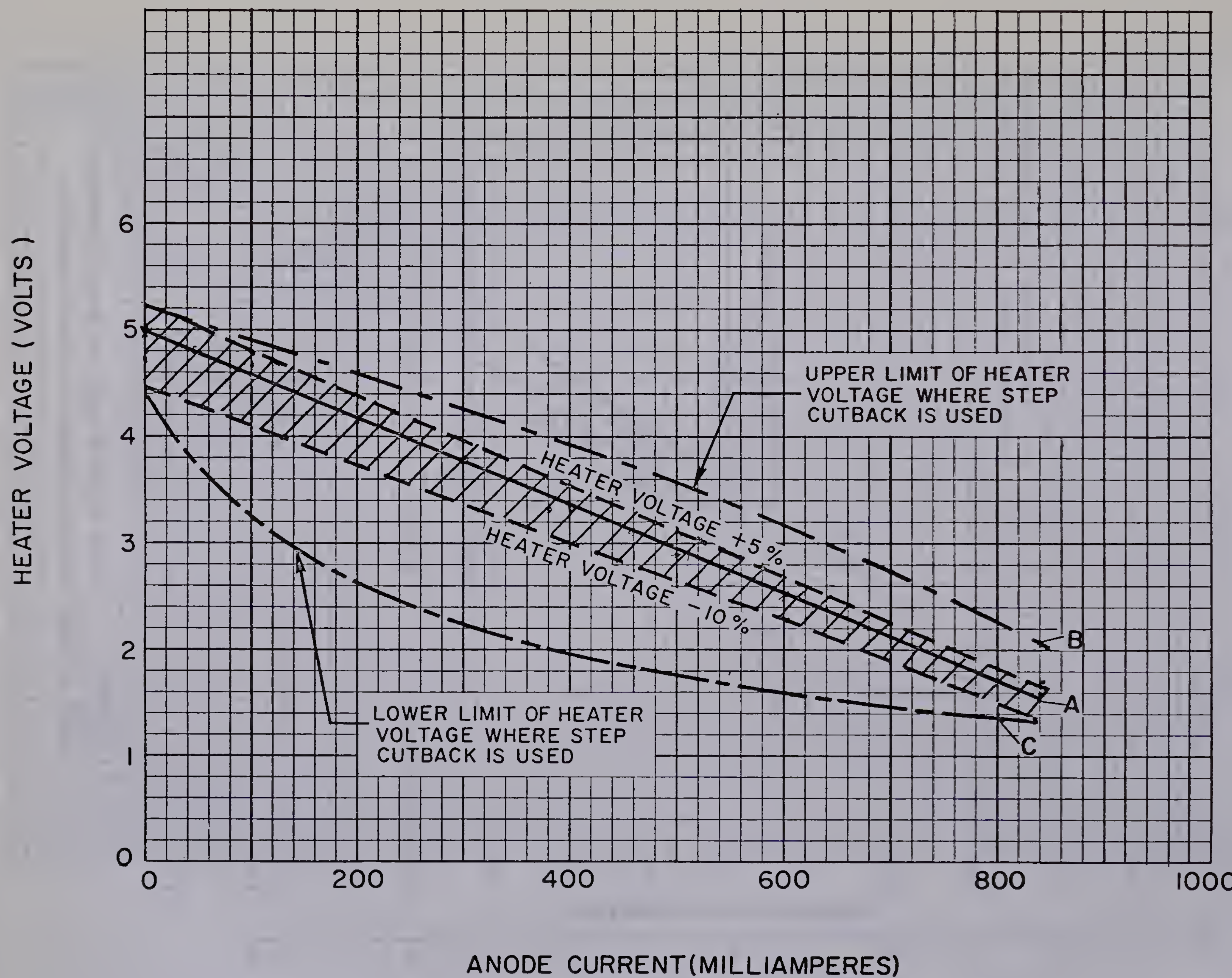

FIGURE 7—HEATER VOLTAGE AS A FUNCTION OF AVERAGE ANODE CURRENT
(SEE OPERATING NOTES)

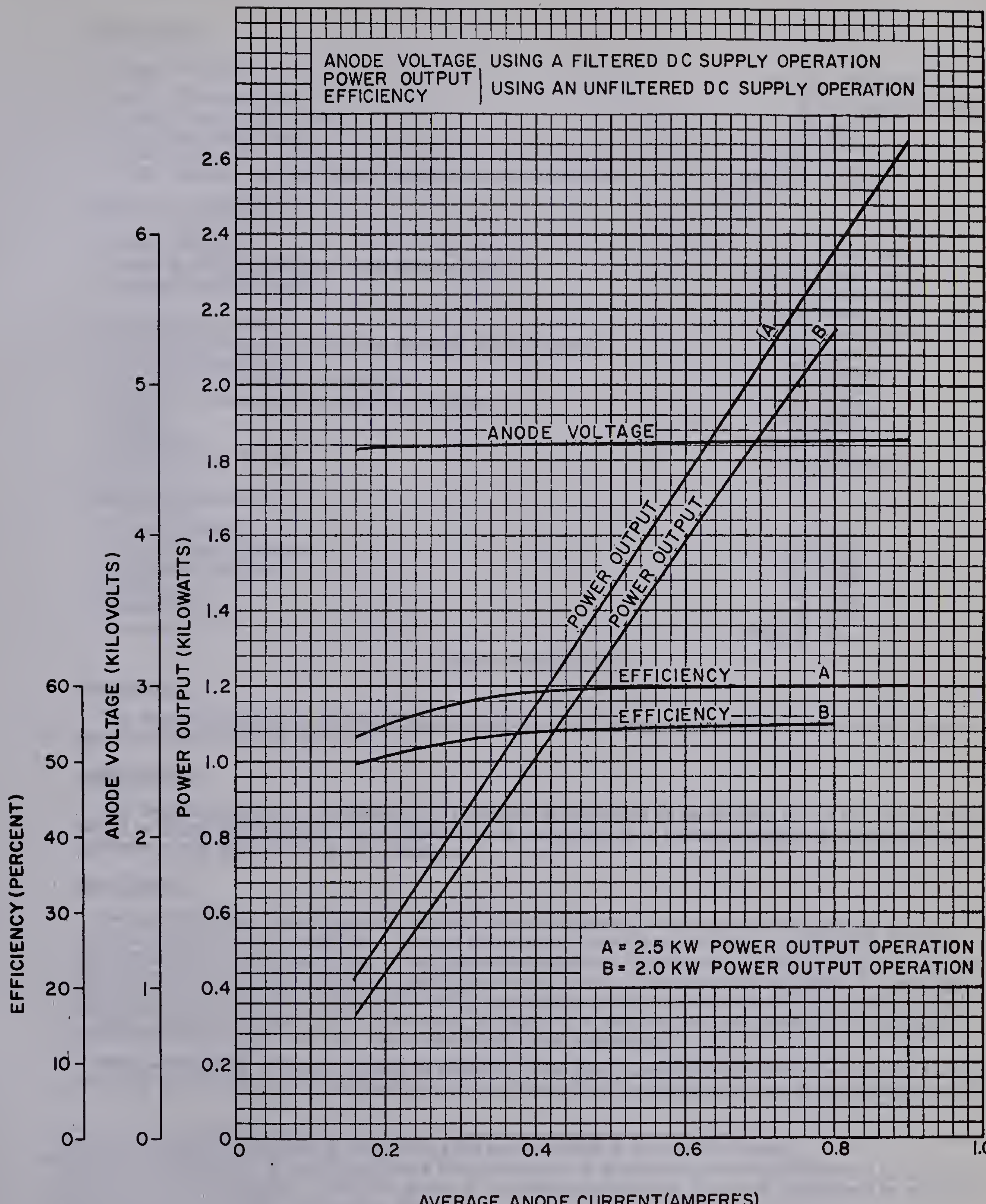

FIGURE 4 – POWER OUTPUT, ANODE VOLTAGE AND EFFICIENCY AS A FUNCTION OF AVERAGE ANODE CURRENT

ELECTRICAL

Heater Voltage[7]	$4.0^{+5\%}_{-10\%}$ volts ac or dc
Heater Warm-up Time[8]	30 seconds minimum
Heater Current at 4.0 volts[9]	30 amps
Cold Heater Resistance	0.018 ohm
Anode Current[10]	0.380 amp
Anode Voltage[5,11] at an Average Anode Current of 0.380 amp	5600±200 volts

ABSOLUTE RATINGS

Heater Voltage	4.2 volts max.
Heater Surge Current (peak value of wave shape)	70 amps max.
Average Anode Current[10]	0.410 amp max.
	0.100 amp min.
Peak Anode Current[12]	1.3 amps max.
Peak Anode Current at 0.380 amp Average Anode Current	0.8 amp min.
Anode Temperature[3]	180°C max.
Heater Input Terminal Temperature	200°C max.
Maximum Temperature at Any Point on Tube[13]	200°C max.
VSWR[1]	4.0 max.
Frequency[5]	2450±25 Mc
Anode-Cathode Voltage	±10 KV max.

TYPICAL OPERATION[1]

Heater Voltage[7]	4.0 volts
Average Anode Current[10]	0.380 amp
Peak Anode Current	1.1 amps
VSWR[1]	3.0 max.
Output Power[5]	1200 watts
Frequency[5]	2450±25 Mc

OPERATING NOTES

Anode Supply

The magnetron should be operated from an unfiltered anode supply with a single phase full wave or two phase half wave rectified system.

Anode Cooling

The magnetron anode is surrounded with a radiator which should be cooled with forced air. The air flow should be ducted to the radiator for efficient cooling and should be of sufficient volume to insure that the maximum anode temperature is not exceeded.

Input Cooling

Because of the high heater current required for the magnetron, it is important that positive input connections be made to the magnetron. This will prevent the contribution of a resistive heating loss to the temperature of the input connectors. This resistance may cause a lower actual heater voltage and result in poor magnetron operation. Therefore, spring type or set screw type connectors should not be used. The Amperex input connectors S-32995 and S-32996 are designed to give the required positive contact and will also aid in cooling the input of the magnetron. Connectors of this design or of a similar clamping type design should be used to make the input connections to the magnetron.

Some of the anode cooling air should be directed on the input connections in order to cool them. A simple way to do this is to mount the tube with the input terminals within the inlet air duct for the anode cooling system.

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[7] The heater standby and operating voltages are the same. There is no cutback schedule.

[8] This time is required at switch-on to insure that the cathode is at proper operating temperature before the application of anode voltage. While the heater is continuously energized, the anode voltage can be switched on and off as desired with no warm-up delay.

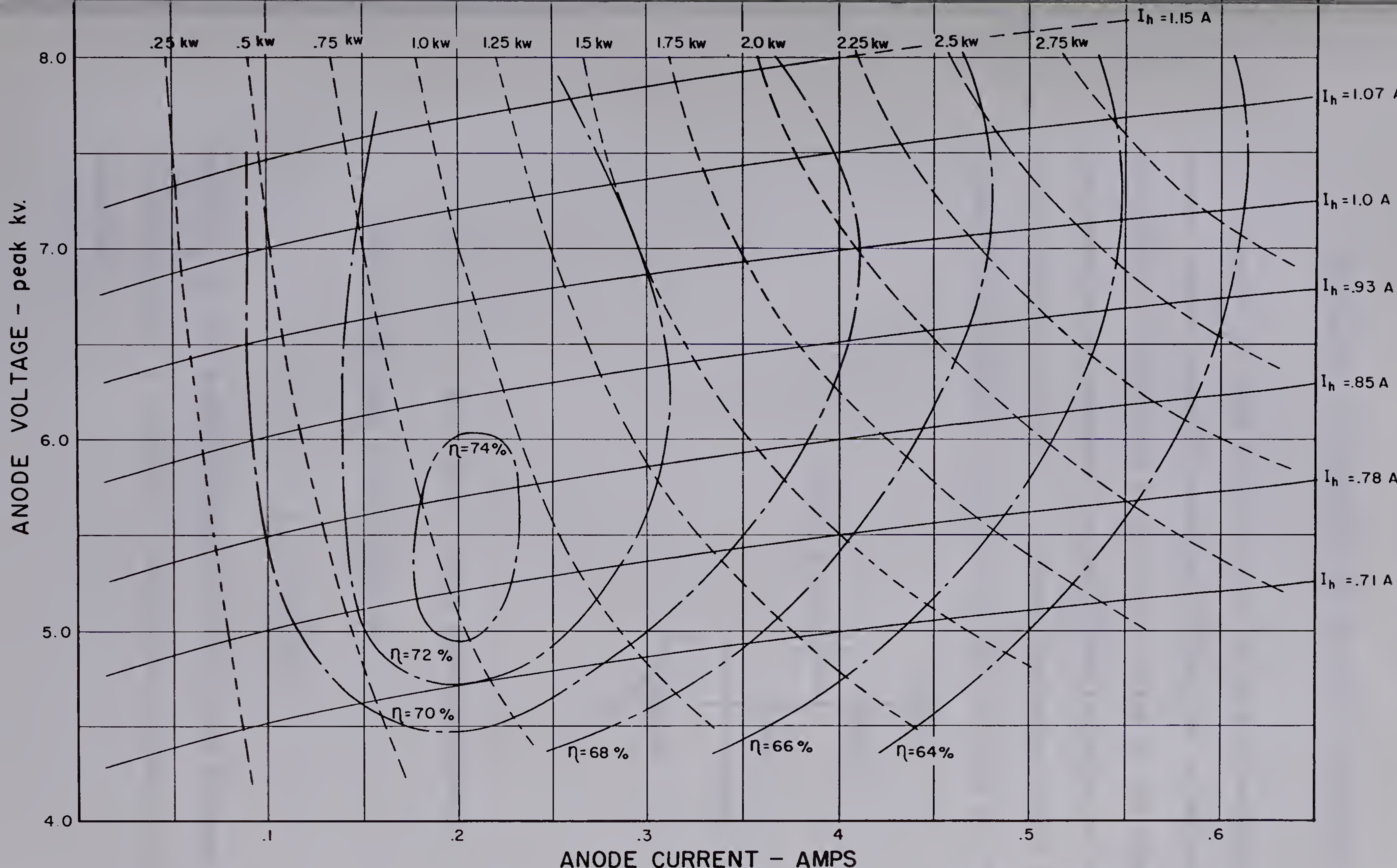
[9] This measurement to be made without the application of anode voltage.

[10] Measured with a moving coil instrument (average value).

[11] Measured with a filtered dc anode supply and with 4.0 volts on the heater. While the anode voltage is measured under the conditions specified, it is not permitted to use the tube in this manner.

[12] The dynamic internal impedance of the anode voltage supply must be adjusted so that the peak anode current is not exceeded during operation.

[13] Measured at the hottest point on the input terminal structure.



L-3821 MAGNETRON PERFORMANCE CHART

CONDITIONS:

(1) MATCHED LOAD
(2) ANODE SUPPLY - FULL WAVE RECTIFICATION (NO FILTER)
(3) $i_{pk} \approx 4 \times I_b$
(4) D.C. FIELD SUPPLY (5 % RIPPLE)
(5) I_h = FIELD CURRENT IN L-3827 MICROTRON UNIT

APPENDIX II

The Effect of Distortion on the R.M.S. Voltage of a Wave

Because the peak distortion in the type of multiplier used in the heater control system is necessarily high, it is desirable to evaluate the effect of distortion on the r.m.s. value of the signal. The following analysis was performed to set allowable limits on the peak distortion of the multiplier circuit. Because distortion occurs primarily in the input diode arrangement, the analysis will be done for this stage.

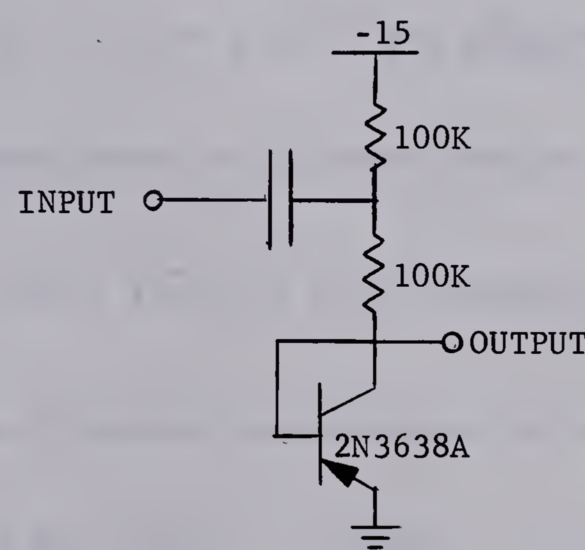
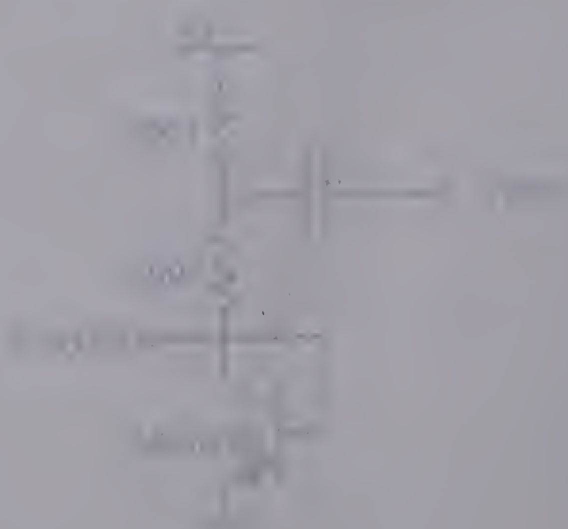


Fig. A2.1 Input Diode Arrangement

The voltage gain of the stage is

$$A_v = \frac{R_d}{100,000} \quad \text{... A2.1}$$

where R_d is the dynamic impedance of the diode expressed in ohms. R_d can be expressed as a function of the current through the diode by the following equation



$$R_d = \frac{26}{I_C + i_c} \text{ ohms.} \qquad \ldots \text{ A2.2}$$

I_C is the quiescent diode current expressed in milliamps and i_c is the signal current in the diode also expressed in milliamps.

The signal current i_c can be expressed in terms of the input voltage v_i which is expressed in volts.

$$i_c = \frac{v_i(1000)}{100{,}000} \text{ milliamps.} \qquad \ldots \text{ A2.3}$$

The voltage gain of the stage is therefore

$$A_v = 2.82 \times 10^{-3} \left[\frac{1}{1 + 0.1082\, v_i}\right]. \qquad \ldots \text{ A2.4}$$

This can be expanded into the following series:

$$A_v = 2.82 \times 10^{-3}[1 + \sum_k (-1)^k (0.1082 v_i)^k] \qquad \ldots \text{ A2.5}$$

Now, for v_i small, the following approximation can be made:

$$A_v = 2.82 \times 10^{-3}(1 - 0.1082 v_i + (0.1082 v_i)^2). \qquad \ldots \text{ A2.6}$$

In order to get a useful error coefficient, the following term is defined:

$$E_{rms} = \frac{\text{RMS value of } V_{oi} - \text{RMS value of } V_o}{\text{RMS value of } V_{oi}} \times 100\% \qquad \ldots \text{ A2.7}$$

V_{oi} is defined as the output from an ideal system which in this case would mean that R_d is independent of the signal current i_c. V_o is the output from the system under consideration.

In order to compare the above error coefficient with a more familiar measurement of distortion, the peak non-linearity is also calculated.

$$E_p = \frac{V_{oip} - V_{op}}{V_{oip}} \times 100\% \qquad \text{... A2.8}$$

V_{oip} is the peak output which would be obtained if the system were ideal, and V_{op} is the output of the system under consideration.

The above error coefficients were calculated for different value of peak input and the results appear in the following table.

TABLE A2.1

V_i	E_p	E_{RMS}
.0923 sin θ	1%	-.00375%
.1846 sin θ	2%	-.015%
.462 sin θ	5%	-.0938%
.923 sin θ	10%	-.375%

APPENDIX III

The Effect of Phase Shift on the Addition of Waveshapes

It is shown in Appendix V that if two waves are identical in shape and in phase, the r.m.s. value of the superimposed waves will be equal to the sum of the r.m.s. values of the individual waves. The effect of adding two waves which are slightly out of phase is considered in this Appendix.

In order to get a worst case analysis, two waves of equal magnitude but out of phase by ϕ degrees are considered summed. Let these waves be $\sin(\theta)$ and $\sin(\theta + \phi)$.

Using the expansion

$$\sin(x + y) = \sin(x)\cos(y) + \cos(x)\sin(y) \qquad \ldots \text{A3.1}$$

$\sin(\theta) + \sin(\theta + \phi)$ can be expressed as $(K_1\sin(\theta) + K_2\cos(\theta)$ where

$$K_1 = 1 + \cos(\phi) \qquad \ldots \text{A3.2}$$

and

$$K_2 = \sin(\phi). \qquad \ldots \text{A3.3}$$

The r.m.s. value of $\sin(\theta) + \sin(\theta + \phi)$ is

$$\sqrt{\frac{1}{\pi} \int_0^\pi K_1^2 \sin^2(\theta) + K_1K_2\sin(\theta)\cos(\theta) + K_2^2 \cos^2(\theta)} \qquad \ldots \text{A3.4}$$

which is equal to

$$\sqrt{\frac{K_1^2 + K_2^2}{2}} \qquad \ldots \text{A3.5}$$

In order to set a limit on the phase shift for the multiplier, equation A3.5 was evaluated for $\phi = 1°$, $2°$, $5°$, and $10°$. The following error was calculated for each value of ϕ. The results are shown in Table A3.1.

$$E_p = \frac{(\sin\theta + \sin\theta)_{RMS} - (\sin\theta + \sin(\theta + \phi))_{RMS}}{(\sin\theta + \sin\theta)_{RMS}} \times 100\% \qquad \ldots \text{A3.6}$$

TABLE A3.1

1°	.00625%
2°	.015%
5°	.0837%
10°	.385%

It should be noted that phase shift is usually associated with a change in gain of a circuit. For example, if the gain was dropping off at the rate of 20 decibels per decade of frequency, a phase shift of 10° would be associated with a drop in gain of 1.48%. This in itself would cause a larger error than the error caused by adding waves which were 10° out of phase.

APPENDIX IV

Calculation of the Transfer Function of the S.C.R. Trigger Circuit

The voltage waveshape associated with S.C.R. control is shown in Fig. A4.1.

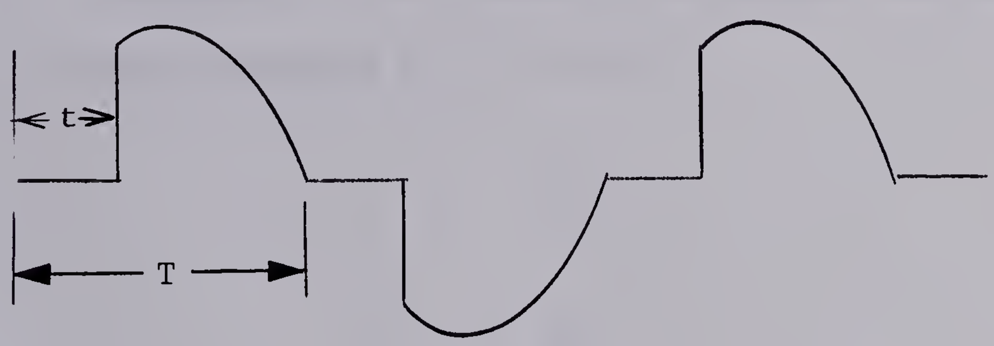


Fig. A4.1 Wave Shape

The r.m.s. voltage is

$$V_{RMS} = V_p \sqrt{\frac{1}{\pi} \int_0^{\left(\frac{T-t}{T}\right)\pi} \sin^2\theta d\theta} \qquad \text{... A4.1}$$

The relaxation oscillator stage of the trigger circuit is shown in Fig. A4.2.

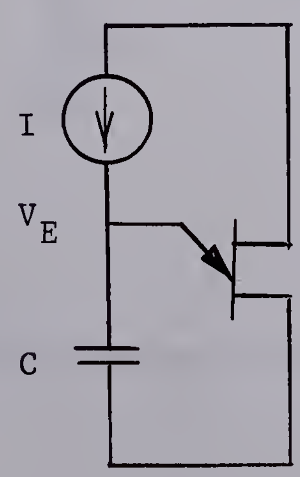


Fig. A4.2 Relaxation Oscillator Stage

The emitter voltage of the unijunction transistor (V_E) can be calculated.

$$V_E = \frac{1}{C} \int_0^t I dt \qquad \text{... A4.2}$$

where t is time expressed in seconds.

Now define V_F as the value of V_E required to fire the unijunction transistor. Assuming that the current I is constant and that t is the time required to charge capacitor C to V_F then

$$V_F = \frac{1}{C} t \qquad \text{... A4.3}$$

and

$$t = \frac{V_F C}{I} \qquad \text{... A4.4}$$

The current I can be calculated in terms of the input voltage V_C to the trigger circuit.

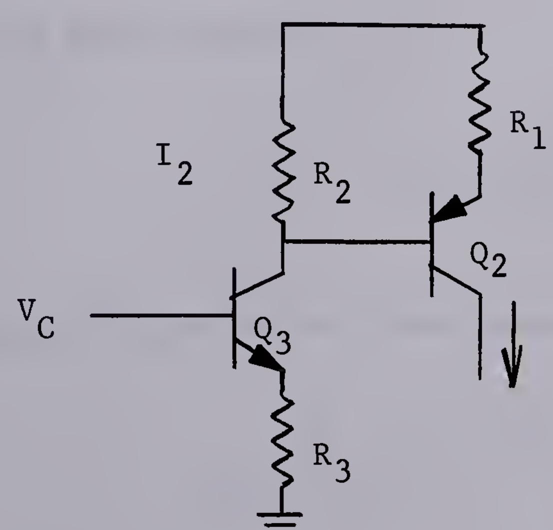


Fig. A4.3 Input Stage of the Trigger Circuit

$$I = \frac{V - V_{BE_2}}{R_1} \qquad \text{... A4.5}$$

but

$$V = I_2R_2 \qquad \text{... A4.6}$$

Therefore:

$$I = \frac{I_2R_2 - V_{BE_2}}{R_1} \quad \text{if } I_2R_2 > V_{BE_2} \qquad \text{... A4.7}$$

The current I_2 can be calculated.

$$I_2 = \frac{V_C - V_{BE_3}}{R_3} \quad V_C > V_{BE_3} \qquad \text{... A4.8}$$

Therefore substituting into equation A4.7

$$I = \frac{\left(\frac{V_C - V_{BE_3}}{R_3}\right)R_2 - V_{BE_2}}{R_1}, \qquad \text{... A4.9}$$

and substituting into equation A4.4

$$t = \frac{V_F CR_1R_3}{(V_C - V_{BE_3})R_2 - R_3V_{BE_2}} \qquad \text{... A4.10}$$

Now substituting this value of t into equation A4.1

$$V_{RMS} = \sqrt{\frac{1}{\pi}\int_0^{K\pi} V_p^2 \sin^2\theta d\theta} \qquad \text{... A4.11}$$

Where

$$K = \frac{\frac{V_F CR_1R_3}{(V_C - V_{BE_3})R_2 - R_3V_{BE_2}}}{T} \qquad \text{... A4.12}$$

This equation was evaluated for conduction angles between 0 and 180 degrees in intervals of 5 degrees. Values of V_C were calculated. The calculations were made assuming the following.

$$V_{BE_2} = V_{BE_3} = 0.65 \text{ volts}$$

$$V_F = 9.0 \text{ volts}$$

$$C = 0.5 \text{ microfarads}$$

$$R_1 = R_2 = 1000 \text{ ohms}$$

$$R_3 = 820 \text{ ohms}$$

$$T = 1/120 \text{ second}$$

The results are shown in the following table. α is the conduction angle expressed in degrees. V_p is chosen so that V_{RMS} will be 1.0 volts for a conduction angle of 180 degrees.

TABLE

α	V_C	V_{RMS}
0	1.62	0.0
5	1.63	0.0119
10	1.65	0.0335
15	1.67	0.0613
20	1.69	0.0939
25	1.71	0.130
30	1.73	0.170
35	1.75	0.212
40	1.78	0.256
45	1.80	0.301
50	1.83	0.348
55	1.86	0.395
60	1.89	0.442
65	1.92	0.489
70	1.97	0.535
75	2.01	0.581
80	2.05	0.625
85	2.10	0.667
90	2.16	0.707
95	2.33	0.745
100	2.40	0.781
105	2.48	0.814
110	2.57	0.845
115	2.68	0.872
120	2.70	0.897

α	V_C	V_{RMS}
125	2.95	0.919
130	3.130	0.938
135	3.340	0.954
140	3.610	0.967
145	3.96	0.977
150	4.43	0.985
155	5.10	0.991
160	6.05	0.996

APPENDIX V

R.M.S. Addition of Voltages

If two waveshapes u(t) and v(t) are added, the r.m.s. value of u(t) + v(t) is not necessarily equal to the sum of the r.m.s. values of the individual waveshapes. Conditions for the r.m.s. addition of waves are established in this appendix. Only waveshapes periodic in period T are considered.

The r.m.s. value of a wave u(t), periodic in period T, is defined as

$$U_{RMS} = \sqrt{\frac{1}{T}\int_0^T u^2(t)dt} \;. \qquad \ldots \text{ A5.1}$$

Similarly, V_{RMS} can be defined as the r.m.s. value of the waveshape v(t).

Now, if W_{RMS} is defined as the r.m.s. value of two superimposed waveshapes u(t) and v(t), then:

$$W_{RMS} = \sqrt{\frac{1}{T}\int_0^T (u(t) + v(t))^2 dt} \qquad \ldots \text{ A5.2}$$

and

$$W_{RMS} = \sqrt{\frac{1}{T}\int_0^T u^2(t)dt + \frac{1}{T}\int_0^T v^2(t)dt + \frac{2}{T}\int_0^T u(t)v(t)dt} \;. \qquad \ldots \text{ A5.3}$$

But:

$$U_{RMS} + V_{RMS} = \sqrt{\frac{1}{T}\int_0^T u^2(t)dt \;+\; \frac{1}{T}\int_0^T v^2(t)dt} \qquad \ldots \text{ A5.4}$$

which is not necessarily equal to W_{RMS}.

There are special cases in which the following equation is satisfied.

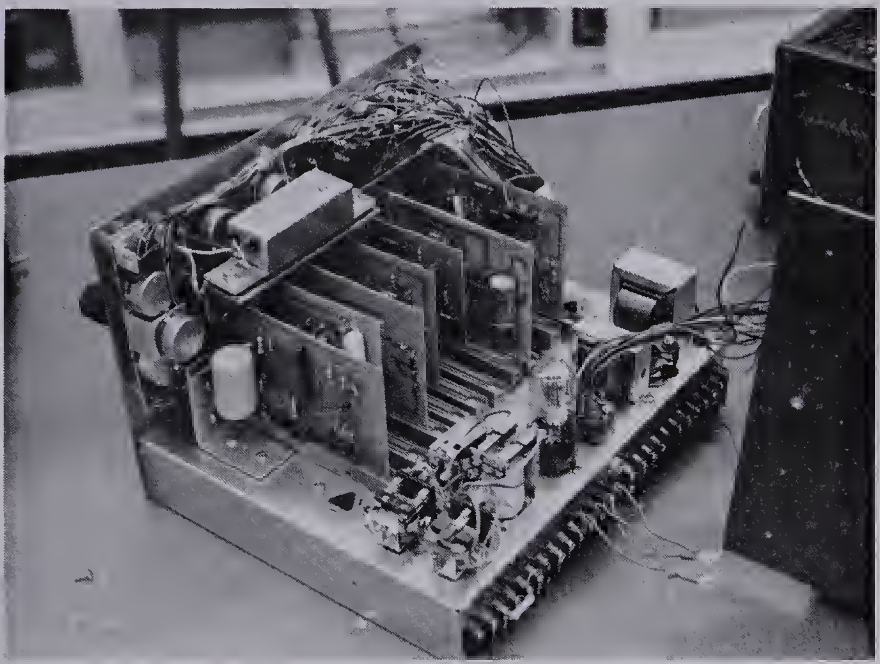
$$\sqrt{\frac{1}{T}\int_0^T (u(t) + v(t))^2 dt} = \sqrt{\frac{1}{T}\int_0^T u^2(t)dt} + \sqrt{\frac{1}{T}\int_0^T v^2(t)dt} \;. \qquad \ldots \text{ A5.5}$$

It can be shown that if $v(t) = Ku(t)$, where K is a positive real number, equation A5.5 will be satisfied.

It can, therefore, be concluded that if two waves are identical in shape (not necessarily magnitude), and in phase, the r.m.s. value of the superimposed waves will be equal to the sum of the r.m.s. values of the individual waves.

APPENDIX VI.

This Appendix contains photographs of the completed supply.



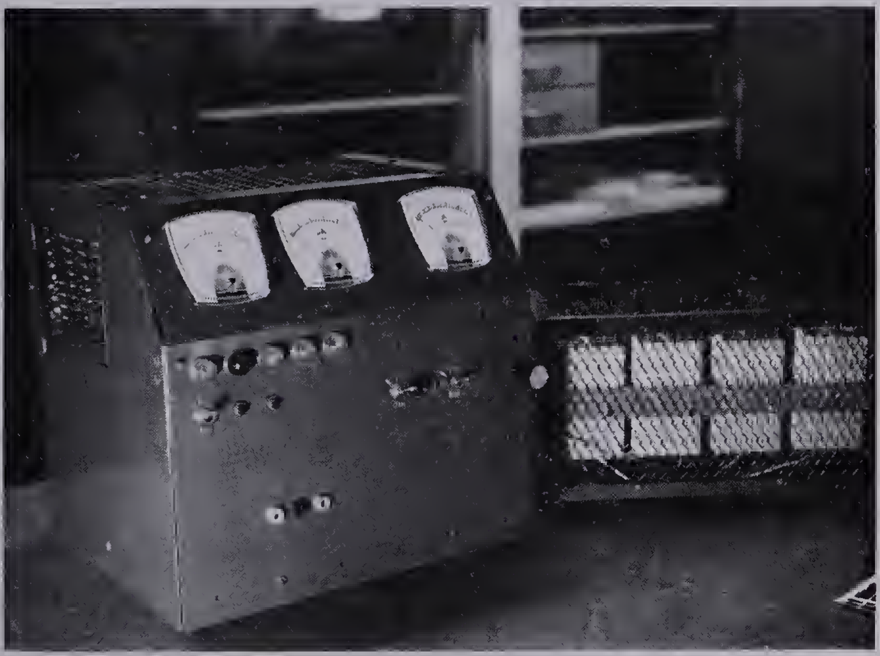


Fig. A6.1   Photographs of the Completed Supply.

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